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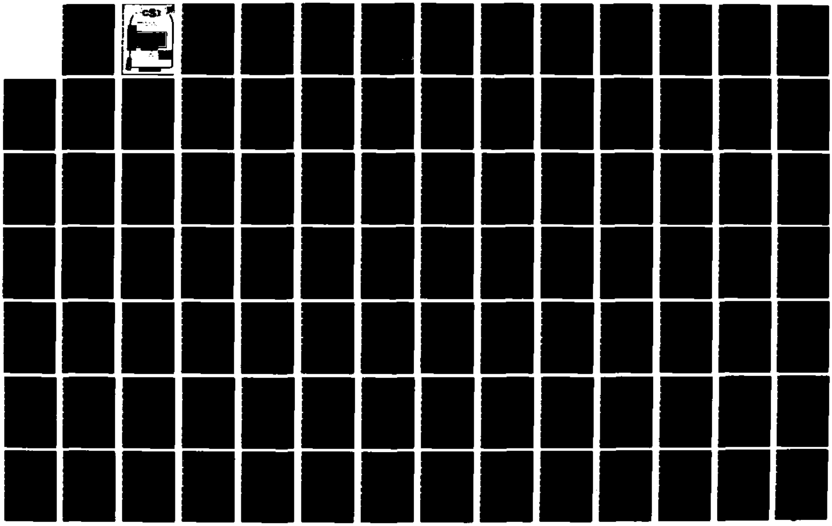

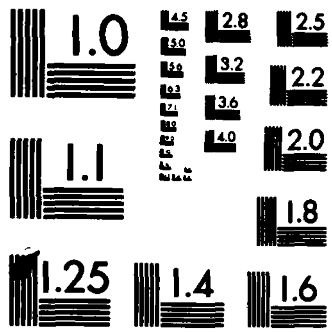

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RESEARCH TRENDS IN MILITARY COMMUNICATIONS

May 1 - 4, 1983

CSI-83-12-01

UNIVERSITY OF SOUTHERN CALIFORNIA
ELECTRICAL ENGINEERING ~ SYSTEMS
LOS ANGELES, CALIFORNIA 90089-0272

RESEARCH TRENDS IN

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20. ABSTRACT (Continue on reverse side if necessary and identify by block number)

This document contains the proceedings of the workshop "Research Trends in Military Communications," held May 1-4, 1983, in Wickenburg, Arizona. Sponsored by the Army Research Office (under Contract DAAG29-83-M-0065) and organized by the Communication Sciences Institute of the University of Southern California, the workshop had as its objective the review of basic research in spread spectrum communications, the evaluation of present needs, and the determination of fruitful future research areas.

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Panels were formed in four basic areas: (a) array signal processing in the spread spectrum environment, (b) spread spectrum communication in jamming, (c) applications of coding to spread spectrum communication, and (d) spread spectrum networks. Efforts were made to balance every panel with participants from each of the academic, industrial, and military laboratory communities. Each panel session consisted of formal presentations by each member followed by at least a one-hour general discussion.

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May 1 - 4, 1983

TABLE OF CONTENTS

PREFACE

This document contains the proceedings of the workshop "Research Trends in Military Communications," held May 1-4, 1983 in Wickenburg, Arizona. Sponsored by the Army Research Office (under Contract DAAG29-83-M-0065) and organized by the Communication Sciences Institute of the University of Southern California, the workshop had as its objective the review of basic research in spread spectrum communications, the evaluation of present needs, and the determination of fruitful future research areas.

Panels were formed in four basic areas: **(a)** array signal processing in the spread spectrum environment, **(b)** spread spectrum communication in jamming, **(c)** applications of coding to spread spectrum communication, and **(d)** spread spectrum networks. Efforts were made to balance every panel with participants from each of the academic, industrial, and military laboratory communities. Each panel session consisted of formal presentations by each member followed by at least a one-hour general discussion.

All sessions were tape recorded and transcribed in an effort to accurately preserve the sense of the discussions. Transcripts of presentations were edited for clarity and for making references to the speaker's slides. The results were approved by the presenters. Hence the following proceedings are, for the most part, not formally prepared papers, but edited transcripts. Some speakers also submitted prepared documents on their area of interest, and these also are included in the proceedings.

Special thanks for the production of these proceedings goes to Mrs. Milly Montenegro, the workshop secretary and typist. Thanks for a job well done also go to John Silvester, Mary Ann Kiefer, and Peter Pawlowski, who along with the workshop chairmen, edited the proceedings manuscript.

R. A. Scholtz

Dr. Robert A. Scholtz

Charles L. Weber

Dr. Charles L. Weber
Workshop Co-chairmen

SESSION 1 – ARRAY SIGNAL PROCESSING IN A SPREAD SPECTRUM ENVIRONMENT

PAUL FEINTUCH

Our first session deals with Array Signal Processing in a Spread Spectrum environment. There exist really two distinct approaches to developing immunity to jamming. One is spatial processing; the other is temporal or signal processing. The spatial processing relies on the jammer and signal being in two distinctly different directions so that it is possible for the adaptive processor to estimate the energy arrival angle of the jammer and adjust its spatial response to put a null in that direction while preserving an almost full main lobe response in the direction of the signal. You can view this as pre-whitening the spatial noise field which contains the jammer as one of its principle components, and then attempting to match the beam peak to the plain wave energy arrival angle of the signal. The temporal approach instead takes the signal, spreads it out in frequency by hopping or modulating with a pseudonoise sequence, and then preserves that information for the receiver to use while denying that information to the jammer. Therefore, the temporal part of the problem can also be viewed as a matched filter.

The primary question in applying Adaptive Array Processing to the Spread Spectrum environment is to what extent these two matched filtering operations interfere with one another. Among the potential problems are those that would be encountered in applying an adaptive array to any problem. The adaptive array needs a reference waveform, or stored spatial replica of the signal to drive the adaptive algorithm. In this case we have to decide whether that reference should be the spread or the despread signal. There is also a convergence process involved in any of the adaptive algorithms. The question arises as to whether or not the adaptive algorithm is going to converge, or synchronize, prior to synchronization by the spread spectrum modem. Then there is the problem of noise being produced by the adaptive algorithms' learning process which can impact the spread spectrum operation. Another consideration is the fact that when you view the spatial null that this adaptive processor is going to place on the jammer, you envision that null at a single frequency. If you change frequencies, the aperture of your array is going to change and the null is going to change in location, or in depth. Yet we need to preserve the nulling properties of the adaptive process over the entire bandwidth of the spread signal if we are going to combine both techniques. So you can see from this that we really have an interdisciplinary problem here. I think most people, as I, have worked in one area or the other, but not in both. What our panelists are going to try to do today is put these two areas in the proper context.

Our first speaker will be Irv Reed. He's going to give us two talks. One is a brief history or an introduction to adaptive arrays. His second talk will come after the break. To inspire the discussion he'll present a whole new algorithm for spread spectrum communications. John Bailey will look at wideband systems and the need to apply sidelobe cancelling to them to preserve the null over the entire spread spectrum band. Marlin Ristenbatt will give us some of his thoughts and concerns on applying the adaptive arrays and the PN sequences. Jim Dupree will present a generally unknown application, namely, the TDRSS satellite system. Bob Dinger has a novel approach to developing the adaptive weights for the whole array. He uses parasitic elements and coupling to set all the weights rather than trying to

determine weightings on each element independently. So with that, our first speaker will be Irv Reed.

IRVING REED

What I've done is prepared a brief history of adaptive arrays as I know the subject. Undoubtedly I left many people out in the history because of the shortness of it, so it is necessarily incomplete. It extends over 26 years.

Brief History of Adaptive Arrays

This is a brief and necessarily incomplete history of adaptive array technology, extending over the past 26 years. The author was a participant in this development and is only human. Thus, as it might be expected, this treatment of the development of adaptive arrays is somewhat biased.

The first practical technique for electronically steering an antenna null in the direction of a jammer was invented in 1957 by the late Paul W. Howells while at the General Electric Corporation, Syracuse, NY. This novel concept was patented, applied for in 1959, awarded 1963, as US Patent #3202990 with title, "Intermediate frequency side-lobe canceller".

My first knowledge of Paul Howells' work in adaptive side-lobe nulling came from a visit to his laboratory in 1957. During this time period at MIT Lincoln Laboratory a method for automatically detecting a train of pulse Doppler radar returns was conceived by Edward J. Kelly and myself. Though it was called by us a "periodogram detector", it was actually an early version of a filter-bank detector for a pulse-Doppler radar. About the same time that Ken Perry implemented the periodogram detector, we at Lincoln Laboratory heard of Paul Howells' VICI (for "velocity-indicating coherent indicator") at G.E., Syracuse.

Although the physical realizations of VICI and the periodogram detector were vastly different, my visit with Paul Howells convinced us that both systems could accomplish identical functions. The groups at GE and Lincoln Laboratory had invented independently optimum matched-filter systems for the detection of Doppler'shifted pulse trains, received from a pulse-Doppler radar. His method was a swept frequency method whereas ours was an actual filter bank. He used an integrating delay line ; we used a hybrid digital storage system.

If Paul Howells was disappointed in discovering that he was not the sole inventor of a Doppler filter-bank detector, it was not discernible to me. Besides demonstrating the very substantial capabilities of VICI against clutter, chaff and weather, he told me about a new technique he had for eliminating jamming as well. This was the concept of his now famous side-lobe canceller.

The world of radar and communications was not ready in the late 1950's for the single side-lobe canceller. He was ahead of his time. It was not until 1962 that Paul Howells and Sidney Applebaum successfully tested a five-loop side-lobe canceller against five jammers, that real interest was generated. This success led them a year later to an important opportunity at the Syracuse University Research Corporation (SURC). At SURC Paul Howells along with Sid Applebaum was given a moderately free hand in his Special Projects Laboratory to investigate the applicability of adaptive techniques to several radar programs. These included the over-the-horizon radar for RADC, ballistic-missile defense radars for ARPA, ABMDA and BMDATC and more recently adaptive cancellation of both clutter and jamming in AEW radars. The latter AEW radar program began in 1972 and was sponsored by NRL, GE and

Technology Service Corporation.

The so-called fully adaptive array concept was conceived by Sid Applebaum and reported in his SURC report TR-66-001, Aug. 1965, titled, "Adaptive Arrays". This report, first available to only workers, in the field was later published as a paper in the IEEE Transactions on Antennas and Propagation in 1976.

My first knowledge of Sid Applebaum's report came from a visit made by Lawrence E. Brennan and myself as representatives of the RAND corp. early in 1965. Larry Brennan and I were looking at that time for new concepts to improve the signal-to-clutter performance of airborne radars. This visit to SURC led ultimately to new clutter cancellation techniques using adaptive arrays which are applicable to AEW radars. These concepts were developed on NAVAIR contracts with the Technology Service Corp. from 1969 to 1972 at which time the adaptive AEW radar program came under the sponsorship of NRL with NRL, GE, SURC and TSC as the principle participants.

The Howells-Applebaum adaptive array technique can be derived from a maximization of the signal-to-noise ratio for a colored noise space-time process. Another independent approach to the adaptive array utilizes the recursive least-squares minimization algorithm which is usually the technique that is applied to communications. The LMS algorithm approach to adaptive arrays and filters was spearheaded by Prof. Bernard Widrow and his students at Stanford University.

Widrow introduced his concepts to adaptivity at the 1960 IRE WESCON with the paper, "Adaptive Switching Circuits" by B. Widrow and M.E. Hoff. In this paper he first develops what he later called the LMS algorithm for adaptive learning, filtering and array processing. Related to adaptive filtering is adaptive pattern recognition. An early paper in this area was, "Pattern-Recognizing Control Systems" by B. Widrow and F.W. Smith and presented at the 1963 Computer and Information Sciences Symposium, Wash., DC. To my knowledge, however, the earliest version of an adaptive filtering system was what is called the RAKE system developed by Robert Price in 1957 for compensating for ionspheric multipath. As you know ionspheric communication usually ends up with 1,2 or 3 multipaths versions of the signal and Robert Price developed one of the earliest adaptive filters to compensate automatically for this multipath problem.

An earlier definitive study on adaptive filters was Widrow's report, "Adaptive Filters I", Stanford University Electronics Laboratory TR-6764-6, Dec. 1966. The first published paper on adaptive arrays in the open literature was, "Adaptive Antenna Systems", by B. Widrow, P.E. Mantey, L.J. Griffiths and B.B. Goode in the IEEE Proceedings 1967. Although this well-known paper came after the 1966 SURC report of S. Applebaum, it was the first paper on adaptive arrays published in English. Remarkably, the written work on adaptive arrays both by Applebaum and Widrow was preceded by a thesis written in France in 1965 by H. Mermoz. This thesis was titled, "Adaptive Filtering and Optimal Utilization of an Antenna", at the Institute Polytechnique, Grenoble, France.

The LMS algorithm in the 1967 paper of Widrow and his coworkers is based on the classical method of steepest descent. Though maximum signal-to-noise (MSN) algorithm of Howells-Applebaum and the LMS algorithm of Widrow were found independently by totally different methods, they are, in fact, very similar. For the known or sure signal the MSN and LMS algorithms both converge to the optimum Weiner solution. One way of thinking about an adaptive algorithm is as a Wiener filter where the filter coefficients are found

automatically. This equivalence was demonstrated in 1972 by Brooks and Reed in IEEE AES-8.

The LMS algorithm for adaptive arrays was developed by two of Widrow's former students. In 1969 L.J. Griffiths developed in IEEE Proc. 57 an adaptive algorithm for wide-band frequency antennas. In 1972, O.L. Frost in IEEE Proc. 60 found an adaptive algorithm for arrays with constraints. Their work has important applications in communications and the passive SONAR problem.

Other contributions to adaptive array for use in passive SONAR were V.C. Anderson (1969), H. Cox (1969), and N.L. Owsley (1969). Owsley wrote the important 1969 report, "A Constrained Gradient Search Methods with Application to Adaptive Sonar Arrays", U.S. Naval Underwater Sound Laboratory, Tech. Doc. #2242-207-63. The paper of Frost and the report of Owsley are closely related, yet complementary.

Closely allied to the LMS algorithm of Widrow and coworkers are the adaptive equalizer algorithms developed by R. Lucky and coworkers at the Bell Systems Telephone Laboratory, beginning in 1965. Again I go back to the RAKE idea of Dr. Price and his algorithm as the earliest adaptive equalizer. Lucky's first paper on the subject was, "Automatic Equalization for Digital Communications", BSTJ, Apr. 1965. An adaptive equalizer is an adaptive filter that attempts to adjust for changing channel characteristics. And that's exactly what the RAKE system did. It slowly changes filter coefficients in such a manner as to compensate for the multipath structure in the ionsphere.

In 1971 Dr. R.T. Compton attacked the problem of acquiring a weak, desired, communication signal in the presence of strong jamming or interference by the use of adaptive arrays. Compton first formulated his concept in "Adaptive Arrays: On Power Equalization with Proportional Control", Ohio State University Quarterly Report 3234-1, Dec. 1971. Compton's power equalization technique for adaptive arrays equalized the power out of the array of the desired signal and the interference to a signal-to-noise level that would permit matched filter acquisition of the desired signal at the array output. An important feature of this technique is that no prior knowledge of signal structure is required. This technique was further generalized and improved in 1971 by C.L. Zahm in IEEE AES-9. Another approach to signal lock-up and acquisition which has application to a wide-band JTIDS-like system was described recently by Brennan and Reed in IEEE AES-18 (1981), which I'll talk about later this morning.

The MSN Howells-Applebaum algorithm with application to adaptive arrays for radar has been developed further by Brennan and his coworkers, including John S. Bailey and myself, at Technology Service Corp and more recently Adaptive Sensors, Inc. from 1970 to the present. This group extended the adaptive array concept to include adaptivity in both the spatial and the time domain, in order to achieve jamming cancellation and clutter rejection simultaneously. In 1978 this work culminated in the paper, "Adaptive Arrays in Airborne MTI Radars", IEEE AP-24, by L.E. Brennan, J.D. Mallet and I.S. Reed.

The Brennan group also showed that fast adaptivity can be achieved by the direct method of adaptive weight computation, the so-called sampled-matrix inversion (SMI) technique, i.e., an open loop adaptive algorithm. An analysis of the number of samples required for the SMI method of adaptivity was given in 1974 in paper, "Rapid Convergence Rate in Adaptive Arrays", IEEE AES-10, by Reed,

Brennan and Mallet. This analysis was based on earlier work in 1963 of the statistician M.R. Goodman in Annuals of Math. Stat. Vol. 34. Goodman's work in turn goes back 50 years to the early statistician, J. Wishart and M.S. Bartlett. It is the probability distribution of Wishart applied to a joint Gaussian process which makes possible an exact analysis of the SMI adaptive array technique.

Both the LMS algorithm of Widrow and the MSN algorithm suffer from a slow convergence when there is a wide spread in the eigenvalues of the steady-state covariance matrix. A number of accelerated gradient techniques are developed to improve this situation. One such technique is the conjugate gradient algorithm for adaptive nulling.

Recently large-scale distributed digital processors have become a reality. This development makes real-time SMI adaptive array processing a real possibility. A promising approach to realize this type of algorithm is to use a cascaded processor such as a Gram-Schmidt network. The convergence time of a cascaded Gram-Schmidt adaptive array processor was shown by Brennan, Mallet and Reed in 1977 to be comparable with the SMI algorithm. Chapter 8 of new book, Adaptive Arrays by Monzingo and Miller contains a summary of these results.

Adaptive array processing has come a long way from Paul Howells' side-lobe canceller. Only now after a quarter of a century have such systems reached actual production and field use. Paul would indeed be proud to know that the 25 to 30 dB improvement in performance against jamming and interference that he promised us those past many years has finally come about.

Slide 1 I want to now just go briefly to a few historical slides. As you can see, Slide 1 is the first page of the patent. It was filed in 1959. It was undoubtedly under some sort of secrecy order so it wasn't patented until 1965. I don't know whether the delay of the patent office or the secrecy order caused the delay. It is interesting to note for those people who are concerned with technology transfer, that by not patenting this earlier, the subject did not grow at all from 1959 to 1965. In other words, the only people working on the subject were at SURC. But immediately after this patent came into being, everybody started to work on the subject. Sidelobe cancellers became something that everybody could use. Also interesting to note is that the French independently discovered it, so holding things back is not necessarily a good idea.

Slide 2 is the actual sidelobe canceller. Number 10 in Fig. 1a could be thought of as a radar antenna; it is just a standard dish. Number 12 in Fig. 1a is an OMNI antenna which has a lower gain. The fact that the OMNI has a low gain compared to the high gain radar antenna says that the OMNI will see the actual radar signals with a much lower SNR. Consequently, the signals don't get cancelled in this process. I will go into the mathematics of this later.

Figure 2 in Slide 3 shows the actual circuit that Paul Howells used. By the way, in those days, such circuits were all realized with tubes in his days.

Slide 4 A better diagram is given in a recent paper that Paul Howells wrote wherein he describes the operation somewhat better. Figure 3 shows what he's trying to do. You can think of the actual original antenna pattern as a sinx over x pattern. The OMNI is superimposed on that. One takes the linear combination of those two patterns and puts a null (forces a sidelobe to be zero at that point), so that the jammer coming in the sidelobe at that angle is nulled out. One creates automatically, a linear combination of the

main beam and the sidelobe in such a manner that the sidelobe is cancelled out. This is done by essentially a nonlinear operation - a mixing operation and a filtering operation.

Slide 5 Another way of looking at this is to take away the carrier frequency so that we are dealing only with complex numbers. Consider Figure 4. The output of the main beam is called B, and x is the output of the auxiliary antenna. Take the conjugate of x and mix it with the filtered output. In other words, we create Y=B-wx. Then this result is filtered.

Slide 6 What one is trying to do is the following. First form y which is the output of the main beam, i.e. y=w*x where x is the output of the OMNI or low gain antenna. Then form the error signal z=w*x. The mean square value of z is given by Eq. 2 on slide 6. There, xx* transpose is what one calls a moment matrix M. (See Eq. 3.) Then complete the square as in Eqs. 4-5 and you will find that the mean square error is lower bounded as in Eq. 6. You will find then that this bound holds only when w is as given in Eq. 7, i.e.

$$w = M^{1}\ \overline{XY}^*$$

The Paul Howells/Applebaum criterion attempts to achieve this weight in a recursive manner.

Slide 7 Figure 5 shows the implementation of the algorithm. You put what is called a steering signal s^* into the amplifiers. The array output is $\sum_{n=1}^{N} q_n V_n$. The space-time version of this as shown in Figure 6. It is a more complicated circuit. Information is taken from both tap delay lines and from different antenna elements and a generalized weight vector is formed based on both of these, using the same generalized criteria as before. The so called MSN algorithm can be obtained in this manner.

In the radar problem or the communication problem where you can separate the signal from the noise, i.e., put a null in the direction of the signal, you have, under the noise alone hypothesis, the probability density given in Eq. 1 on Slide 9, assuming Gaussianity. For signal-plus-noise you would get the density given in Eq. 2. If you take the likelihood ratio, as in Eq. 3 you can prove that L(x) is a monotonic function of the quantity $s^*M^{-1}x$. From this, the optimum weights are found to be $W=M^{-1}s$ as in Eq. 6. Finally form the filter $z=w^*x$ with these weights. This is the optimum space time filter.

Slide 10 The array output is given by $\sum_{k=1}^{K} w^*_k x_k = w^*x$. The SNR at the output of the array is given by

$$\frac{S}{N} = \frac{a^2|w^*s|}{w^*Mw}$$

The SNR is maximum when $w=M^{-1}s$. This expression is the generalized SNR ratio for any weight vector.

Slide 11 Now suppose that you want an estimate of the covariance function. You take the estimate $\hat{M}$ of the covariance function M. Next take its inverse. Then $\hat{w} = \hat{M}^{-1}s$ is an estimate of weight vector w. The output SNR is then given by Eq. 2 on slide 11. You can form the ratio of the output SNR to the maximum output SNR which occurs for optimum weights. Call this ratio $\rho(N,k)$ as in Eq. 3. One can find the probability density of ρ. This can be accomplished with the Wishart distribution. You will find the pdf is the Beta function given in Eq. 5. An important quantity one can get out of this is the mean number of samples required, which is given in Eq. 6. This formula means that if you suppose that the number of degrees of freedom is n, and if you have roughly 2n samples of data to estimate the covariance matrix,

then you will lose on the order of only about 3dB sensitivity on the average.

<u>Slide 12</u> The general processor of today looks like the one shown in Figure 7. You have a lot of A-D converters on chips with N signals coming in. You have complex signals I and Q, although this could be done on a low frequency or IF carrier. Then you have a digital processor or some type of high-speed pipe-line processor which now forms the output of the array.

Aug. 24, 1965 P. W. HOWELLS 3,202,990

INTERMEDIATE FREQUENCY SIDE-LOBE CANCELLER

Filed May 4, 1959 2 Sheets-Sheet 1

FIG.1a.

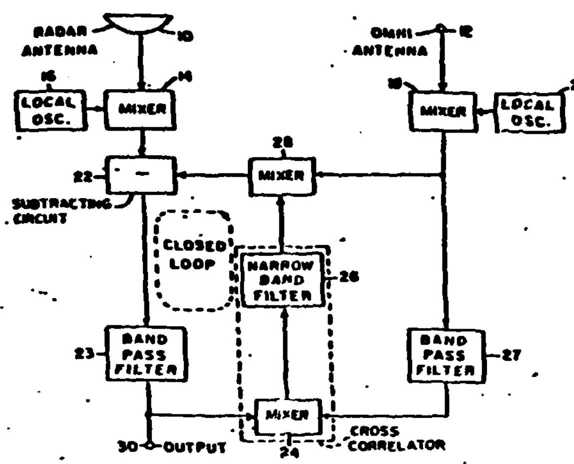
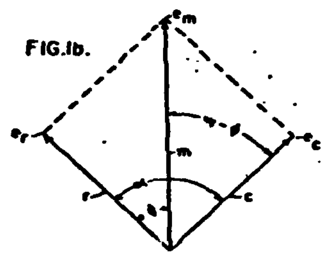


FIG.1b.



SLIDE 2

INVENTOR:
PAUL W. HOWELLS,

BY T. E. Kristofferson
HIS ATTORNEY.

SLIDE 1

United States Patent Office

3,202,990
Patented Aug. 24, 1965

1

3,202,990
INTERMEDIATE FREQUENCY SIDE-LOBE CANCELLER
Paul W. Howells, Morrisville, N.Y., assignor to General Electric Company, a corporation of New York
Filed May 4, 1959, Ser. No. 810,761
9 Claims. (Cl. 343—100)

This invention relates to the elimination of jamming in a signal processing system and more particularly to the provision in a signal reception system of a side-lobe canceller employing a minimal amount of duplication of signal processing equipment.

Antenna side-lobes make it possible for a single jammer to be effective against a radar from any angle of azimuth. The jammer power density at the radar antenna is inversely proportional to the square of the distance between the jammer and the radar. Radar signal power density is inversely proportional to the fourth power of the distance between the target and the radar. The jammer thus has a power advantage that overcomes the radar antenna's reduced gain as its side-lobes scan past the jammer.

Side-lobe cancellation is a fundamental approach to elimination of jamming from received signals. A side-lobe canceller uses jammer location to eliminate jamming signals and in effect maintains a null in the radar antenna receiver pattern in the direction of the jammer. To accomplish this effect the side-lobe canceller uses an omni-directional antenna adjacent to the radar antenna.

Prior art side-lobe cancelling systems have used two separate receivers, one for each antenna, adjusting one in amplitude so their outputs may be subtracted to eliminate the jammer at the video level. This approach has disadvantages since it involves the complex problem of matching two complete receivers almost perfectly in band pass, time delay and amplitude response. Video side-lobe cancellation also requires complete duplication of all intermediate frequency signal processing equipment being used with the radar. Accordingly it is an object of this invention to provide side-lobe cancellation without requiring matching of complete receivers.

Another object of this invention is to provide an improved jammer cancelling signal processing circuit.

Still another object of this invention is to permit intermediate frequency signal processing equipment to be used after side-lobe cancellation has taken place.

Still another object of the invention is to provide flexibility in a side-lobe canceller which can be expanded to cancel multiple jammers.

A further object of this invention is to employ phase information for jammer cancellation as well as the amplitude information normally used in video cancellers.

In carrying out the invention in one form thereof a main signal transmission channel is provided having a directional antenna adapted to receive desired signals. This channel is coupled through a conventional first detector to produce an intermediate frequency signal and through a signal combining circuit and a first filter network to an output terminal. An auxiliary signal transmission channel is provided having an omni-directional antenna adapted primarily to receive undesired signals. The omni-antenna is coupled through a conventional first detector to produce an intermediate frequency signal offset from the intermediate frequency signal in the main

2

transmission channel by a fixed amount at least equal to the signal bandwidth. It is then coupled through a filter network having the same time delay as the filter network in the main channel. The output of the main channel and the filter network in the auxiliary channel are coupled to a band pass cross correlator which has an output signal representing the amplitude and phase difference of the components of the undesired signal in each of the two channels. The cross correlator output is then heterodyned with the offset intermediate frequency signal of the auxiliary channel in order to produce an auxiliary channel output at the same frequency as the intermediate frequency in the main channel and with any phase difference removed. This signal is then coupled to the signal combining circuit in the main channel which subtracts it from the signal in the main channel, closing the loop and providing continuous cancellation of the jamming signal in the output of the main channel.

In another form the invention may employ a number of omni-directional antennas in order to provide for cancellation from several jamming sources.

The novel features which are believed to be characteristic of the invention are set forth with particularity in the appended claims. The invention itself however, together with further objects and advantages thereof can best be understood by reference to the following description taken in connection with the accompanying drawings, in which FIG. 1(a) is a block diagram illustrating one embodiment of the invention employing a single omni-antenna,

FIG. 1(b) is vector diagram showing certain voltage relationship useful to explain the operation of the circuit of FIG. 1(a),

FIG. 2 illustrates in block diagram form another embodiment of the invention extended to be effective against multiple jammers,

FIG. 3 represents waveforms useful to explain the operation of the circuit of FIG. 2.

Referring now to the drawings, in FIG. 1 there is illustrated an embodiment of the invention employing a directional radar antenna 10, and a single omni-directional radar antenna 12. The directional antenna receives a desired signal in its main lobe. Jamming is sometimes introduced by means of a high power source of jamming signals located a distance from the target in order to swamp out target information. This jamming may be picked up by a side-lobe of the radar antenna. In order to properly detect a target it is necessary to eliminate that portion of the signal received by the radar antenna which is attributable to jamming. The output of directional antenna 10 is mixed in a mixer 14 with the output of a local oscillator 16. Mixer 14 can be any conventional first detector. The output of omni-directional antenna 12 is mixed in another conventional first detector 18 with the output of a second local oscillator 20, down to an intermediate frequency which is offset in frequency from the intermediate frequency of the radar jamming signal. It is obvious that other techniques of obtaining the desired intermediate frequency may be employed such as the use of a single local oscillator for both channels and then offsetting the frequency of the two channels. The output of mixer 14 is connected through a subtraction circuit 22 and through a band pass filter 23 to one input of a band pass cross correlator comprising a mixer 24 and a narrow band filter 26. The output of mixer 18

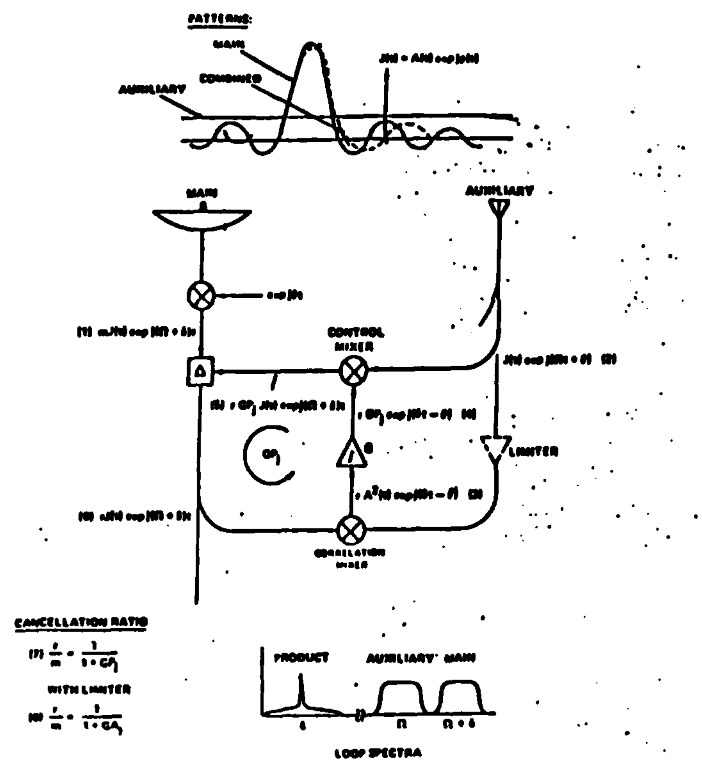


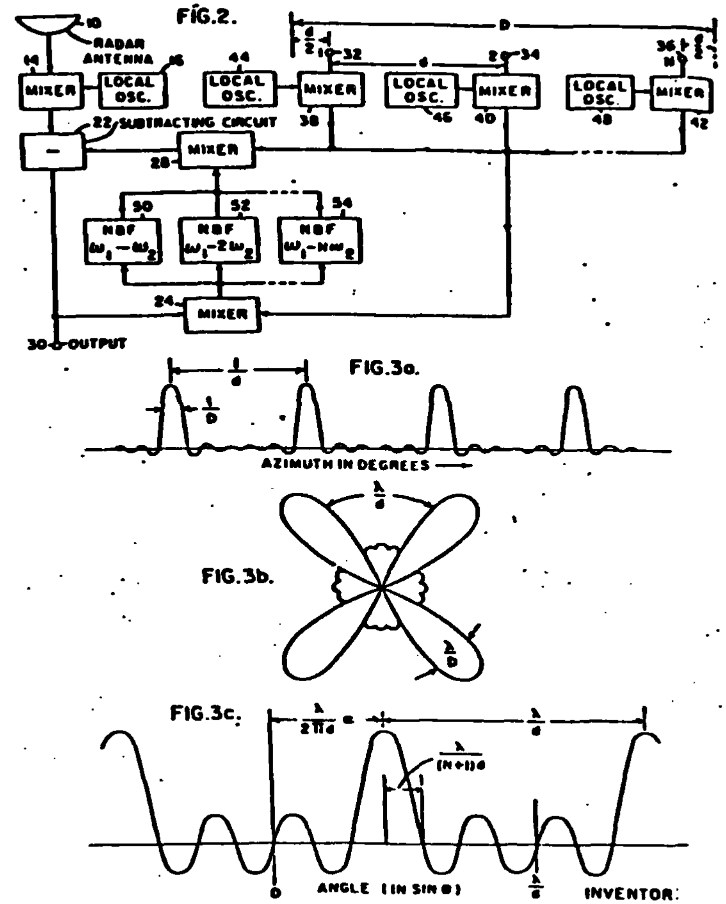
Fig. 1. Correlation loop.

SLIDE 4 Figure 3

Aug. 24, 1965 P. W. HOWELLS 3,202,990

INTERMEDIATE FREQUENCY SIDE-LOBE CANCELLER

Filed May 4, 1959 2 Sheets-Sheet 2



INVENTOR:
PAUL W. HOWELLS,
BY T. C. Kintzinger
HIS ATTORNEY.

SLIDE 3

SIDELOBE CANCELLER

$$Z = Y - W^*X \qquad \text{Eq. (1)}$$

$$\overline{|Z|^2} = \overline{|Y|^2} - W^*\overline{XY^*} - \overline{YX^*}W + W^*\overline{XX^*}W \qquad (2)$$

$$\overline{XX^*} = M \qquad (3)$$

$$\overline{|Z|^2} = \overline{|Y|^2} - \overline{YX^*}M^{-1}\overline{XY^*} + (\overline{YX^*}\,M^{-1}\,\overline{XY^*} - W^*\overline{XY^*} - \overline{YX^*}W + W^*MW) \qquad (4)$$

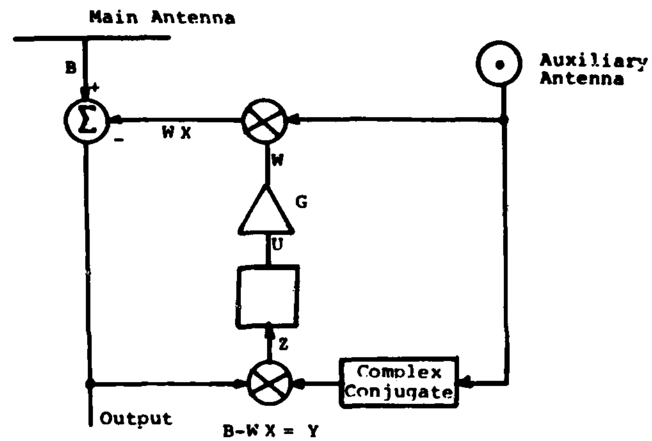
$$= \overline{|Y|^2} - \overline{YX^*}M^{-1}\overline{XY^*} + |M^{-\frac{1}{2}}\,\overline{XY^*} - M^{\frac{1}{2}}\,W|^2 \qquad (5)$$

$$\geq \overline{|Y|^2} - \overline{YX^*}\,M^{-1}\,\overline{XY^*} \qquad (6)$$

$$\text{Equal when:}\quad W = M^{-1}\,\overline{XY^*} \qquad (7)$$

SLIDE 6

ANALOG SIDELOBE CANCELLER - SINGLE AUXILIARY



$$W = GU$$

$$\tau\dot{U} = Z = \frac{\tau}{G}\dot{W}$$

$$Z = X^*(B-XW) = \overline{X^*B} - MW$$

$$M = \overline{XX^*}$$

$$\frac{\tau}{G}\dot{W} + MW = \overline{X^*B}$$

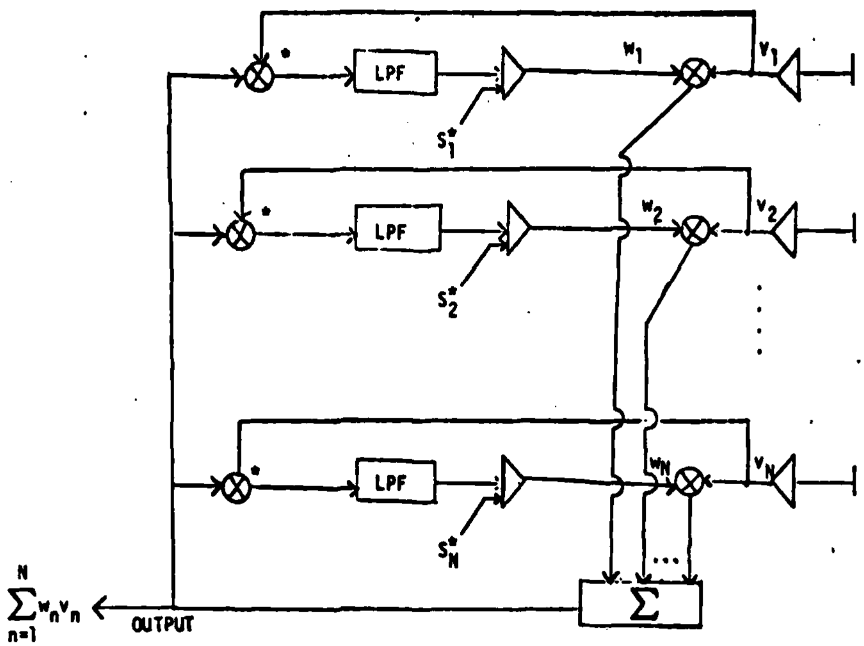
Steady state weight ($\dot{W} = 0$) is

$$MW = \overline{X^*B}$$

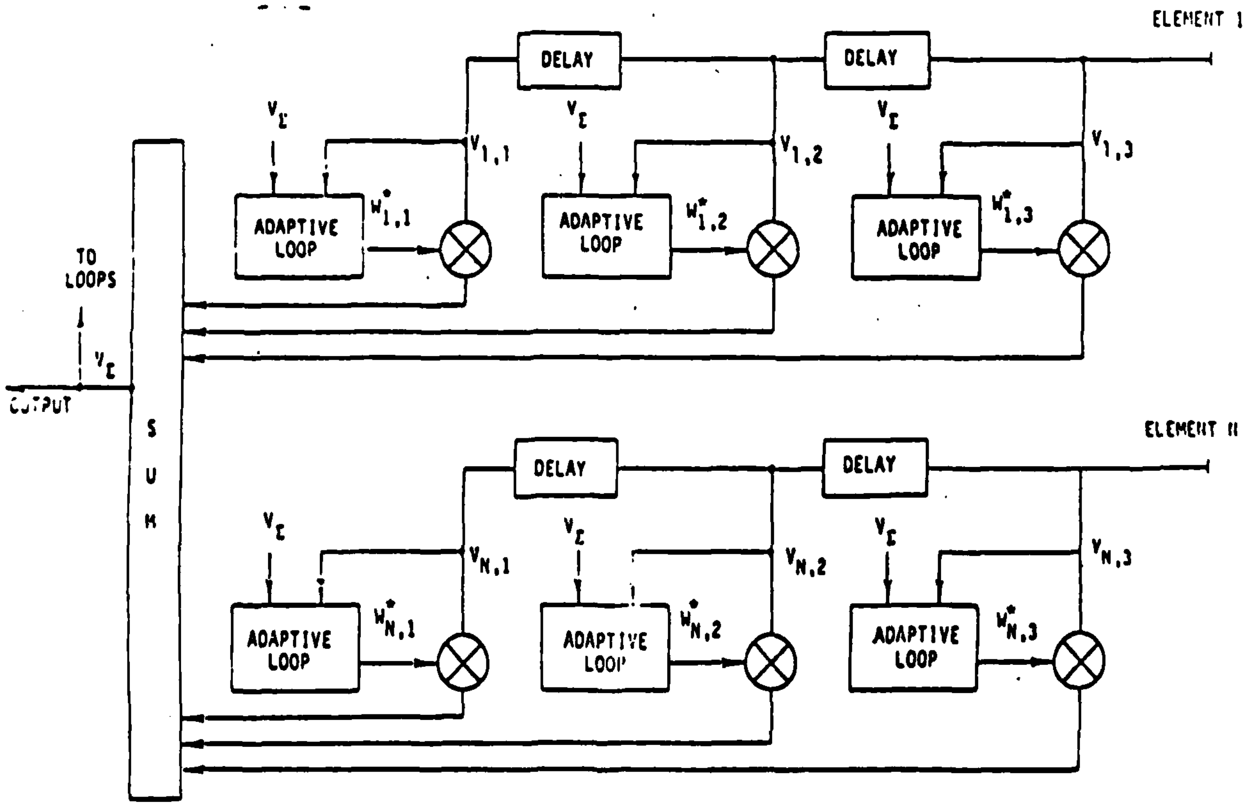
$$W = \frac{\overline{X^*B}}{\overline{X^*X}} \quad \text{optimum weight}$$

Figure 4 SLIDE 5

MAXIMUM S/N ARRAY (Applebaum) -- SPECIFY DESIRED SCAN DIRECTIONS



SLIDE 7 Figure 5



ADAPTIVE ARRAY/TIME PROCESSING

SLIDE 8 Figure 6

$$P_0(X) = \frac{1}{\pi^k|M|} \exp(-X^*M^{-1}X) \qquad \text{Eq. (1)}$$

$$P_1(X) = \frac{}{\pi^k|M|} \exp(-(X^*-S_1^*)M^{-1}(X-S_1)) \qquad (2)$$

$$L(X) = \frac{P_1(X)}{P_0(X)} = \exp(-S_1^*M^{-1}S_1 + S_1^*M^{-1}X + X^*M^{-1}S_1) \qquad (3)$$

$$S_1 = ae^{i\phi}S \qquad (4)$$

$$M = E(NN^*) \qquad (5)$$

L(X) is monotonic increasing function of $S^*M^{-1}X$

Optimum weights α $W^* = S^*M^{-1}$

$$W = M^{-1}S \qquad (6)$$

Output : $Z = W^*X$; compare $|Z|$ with threshold (7)

SLIDE 9

ARRAY OUTPUT $\sum_{k=1}^{K} w_k^* x_k = W^*X$ (Eq. 1)

W = COLUMN VECTOR OF ARRAY WEIGHTS (COMPLEX)

X = COLUMN VECTOR OF ARRAY ELEMENT OUTPUTS (COMPLEX)

OUTPUT SIGNAL POWER = $a^2|W^*S|^2$

S = COLUMN VECTOR OF SIGNAL PHASORS

NOISE POWER = $E|W^*N|^2 = W^*MW$

N = COLUMN VECTOR OF RANDOM NOISE COMPONENTS

M = COVARIANCE MATRIX OF NOISE $\overline{NN^*}$

$$\frac{S}{N} = \frac{a^2|W^*S|^2}{W^*MW}$$

IS MAXIMUM FOR $W \propto M^{-1}S$

ALSO OPTIMUM WEIGHTS FOR SIGNAL DETECTION IN GAUSSIAN NOISE

SLIDE 10

DIGITAL ADAPTIVE ARRAY - APPLY WEIGHTS TO NEW SAMPLES

$$\hat{W} = \hat{M}^{-1} S \qquad \text{Eq. (1)}$$

R_1 = output signal-to-noise ratio

$$= \frac{|\hat{W}^* S|^2}{(\hat{W}^* M \hat{W})} \qquad (2)$$

$$\rho(N,K) = \frac{R_1(N,K)}{R_0(N,K)} \; ; \text{ K elements, N samples} \qquad (3)$$

$R_0(N,K)$ = output signal-to-noise ratio with optimum weights

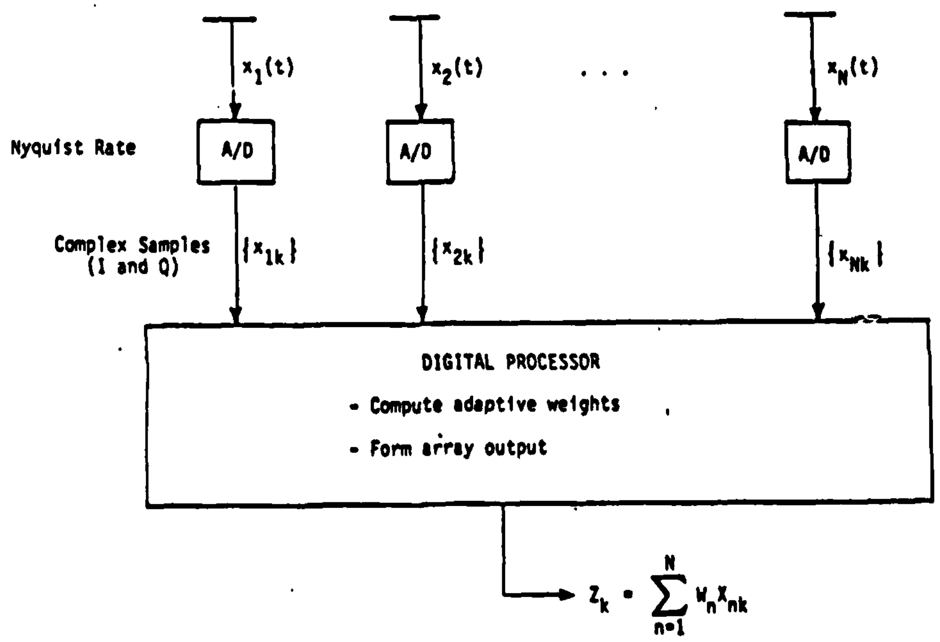
$$= \frac{|W_0^* S|^2}{(W_0^* M W_0)} = S^* M^{-1} S \quad ; \quad W_0 = M^{-1} S \qquad (4)$$

$$P(\rho) = \frac{N!}{(K-2)!(N+1-K)!} (1-\rho)^{K-2} \rho^{N+1-K} \qquad 0 \le \rho \le 1 \qquad (5)$$

$$\bar{\rho} = \frac{N-K+2}{N+1} \qquad (6)$$

SLIDE 11.

DIGITAL ADAPTIVE ARRAY



- Very rapid convergence
- Apply weights to new samples or "same" samples
- Demonstrated experimentally

SLIDE 12 Figure 7.

JOHN BAILEY: I was asked to give an overview of sidelobe cancelling techniques used in wideband systems. Slide 1 I am going to talk about a technique which I think is the wave of the future. I will also be talking about receiver equalization because in wideband systems that becomes a very important consideration.

Slide 2 indicates what we mean by a sidelobe cancellor. The E is a column vector of all the inputs in the system. B refers to a formed beam from elements in an array. See Slide 2 E1 through E4 form auxiliary channels from which one could potentially cancel 4 jammers, that in general would be an N vector. The steady state covariance matrix is the expected value of the outer product where each term of the covariance matrix would be the cross-correlation between the average value of the appropriate terms in the steady state on the assumption of stationary statistics.

What one normally means by sidelobe cancellor is that the array is not fully adaptive. That is, when you apply a steering vector to form an adaptive weight vector you apply a 1 in the direction of the formed beam. In general, however, in a fully adaptive array the steering vector that would be applied adaptively would incorporate the apriori known phase and/or relative amplitude of the signal coming in from a given direction.

The equation $\bar{w} = M^{-1}\bar{s}$ is nothing more than matrix notation for a set of linear equations. All algorithms, either implicitly or explicitly solve this set of linear equations. They could do so by physically inverting the covariance matrix, M, which could be done iteratively as in all the old techniques. I'll get into that shortly, when we go into digital systems.

I'll use an example that doesn't have any mathematics in it but is analogous to the diagram that Irving Reed put up. Take an example of 1 beam and 1 auxiliary channel as shown in Figure 3. Of course this could also just be two OMNI channels in a communication system. What one is trying to do is take the E channel, which is an OMNI, and adjust the relative amplitude and phase, such that when you add it to the beam in amplitude and phase in the direction of the sidelobes, the formed beam will cancel out the jammer. In point of fact, the algorithm we alluded to will maximize the total signal to jamming plus noise ratio in the system. If you form the two by two covariance matrix you'll have 4 terms which will be the expected value of the power in the beam and the auxiliary channel in the cross-correlation terms. One of the very important properties of this incidentally is that the adaptive beam output is orthogonal to the E channel that we are adapting with. More about that later.

Slide 4 The other introductory type of information I'd like to talk about before getting to SLC is just a brief review of the techniques that are currently being used. The old analog systems where we talk about Applebaum/Howell loops, Widrow's algorithm, modern versions that Griffiths and people like Frost more recently proposed, basically take the same set of linear equations and attempt to solve them in some iterative sense, usually with an analog system. The Applebaum/Howell is an algorithm that maximizes signal noise ratio. The Widrow algorithm finds the least mean square residue between the output of the array and some apriori known signal generated as a pilot signal and generated internally. Such things as equalization networks fit into this category. A property of all of these iterative techniques is that the speed of convergence is a function of the structure of the noise field. Specifically the convergence depends upon the eigenvalues of the covariance matrix. Even

though one never physically forms a covariance matrix, they implicitly solve that same set of linear equations. Furthermore, you can show that if the eigenvalues are significantly different, which might correspond to an example of two unequal power jammers, that there does not exist a conceptual implementation that converges *rapidly*. That is, it may take literally thousands of samples of instantaneous bandwidth to make the algorithm converge. Now, as we evolve towards digital technology, we have shown that if instead of using the steady state covariance matrix in the optimum weight, one uses a sample version of that covariance matrix with order 2N samples, you'd be nearly as well off as you would have been in the steady state case. This to me is one of the most fundamentally important results that is now currently being used, and will be used in the future because it will speed up convergence times by orders of magnitude. Just to give an example suppose one had a 5 mHz bandwidth system, corresponding to a 200 nanoseconds narrow pulse with 5 adaptive degrees of freedom in the system. Then in effect that says that after only 10 samples subtending 2 microseconds, one has sufficient information to give a nearly optimum solution as one would have in the steady state analysis. Given that the real world or jamming field may be time variant, for example, due to blinking jammers or whatever, this offers a potential order of magnitude increase in convergence times.

My K is the number of adaptive degrees of freedom, Reed's K was the total number of channels in the system. R is the output signal to residue, where residue is receiver noise plus jamming residue, given that you have K samples of the sample covariance matrix over what would have been achievable with an infinite number of samples. The appropriate probability density distribution is the f-distribution. Once you do these things digitally, you could also potentially form a covariance matrix, let's say with 2N samples, invert it, determine an adaptive weight, and then rather than applying that adaptive weight to new data, you could apply it to the same data. The virtue of that is that this will still work essentially perfectly against any conceivable blinking strategy of time varying noise field. There will no longer exist, for example, blinking strategy of jammers that could defeat that type of system. Essentially you would always be applying the data to what was measured for that same data. What this says is that the measure of performance is essentially the same as it is for reprocessing new data. It takes exactly one sample more when you reprocess the same data. Implicit in what I said is that this achievable performance is now completely independent of the structure of the noise fields. It is no longer eigenvalue dependent. I bring this up in this detail now because it is very important when considering digital configuration for spread spectrum applications. With that background, we'll now delve into an attempt to quantify the widebandwidth problem.

Visualize an array of N elements with jamming coming in some direction of θ. See Slide 5. The distance between the end of the array and the i^{th} filament is d_i. If you write down a formed beam with any set of weights (as in Eq. 1, Slide 5) whether adaptive or deterministic, you'll end up with a beam pattern with the following property: if you switch over to another frequency, since you have f sin θ in the phase term, a change in frequency is equivalent to a change in sine of angle space and therefore you will have different sidelobes structures at different frequencies. Therefore, in a wide bandwidth system it will be impossible in principle to perfectly cancel jammers at all frequencies because they have different

patterns. The useful artifice for looking at this will take advantage of a fairly intuitive algorithm. Slide 6 If you have a broadband system with a point source, as shown in Slide 6, with different patterns at different frequencies, it mathematically equivalent to a narrowband system 1 frequency where the jamming source is spatially spread. In this case, the shape of the beam would be the receivers response appropriately scaled in sine of angle space.

There are two different ways of looking at the broadband problem. One can look Slide 7 in the time-domain as in Slide 7 and instead of bandwidth limiting, you time-limit the system. You can think of a narrow pulse of duration T, and its bandwidth is 1/T. If there's a time delay, and we're trying to use E2 to cancel E1, then literally there's nothing there for part of the pulse. Therefore, no matter how well you do, there will be jamming residue. I point this out because now when we talk about various ways in which you can cope with such wideband systems, this will suggest techniques.

Slide 8 Eq. 2 on Slide 8 represents the actual spatial spread that you would expect in a wideband system in absolute terms. The d-θ/df in Eq. 1 is the frequency sensitivity due to the multiplication of the frequency times and sine of the angle. One useful way to say how much that bothers you is to ask, well how much does that effect spatial spread in beamwidths. So if you normalize $\Delta\theta$ by the matched filter beamwidth in the same direction, you end up with the ratio given in Eq. 3. Coincidently, you can also think in terms of time-bandwidth product which signal processing types might prefer. By time-bandwidth product I mean the time delay in free space across the aperture which would be

$$TB = \frac{A \sin \theta}{c} B$$

as shown in Eq. 4. Now what can be shown is that if you have a full aperture, and you ask what is the achievable cancellation ratio in wide bandwidth system, you end up with Eq. 5. This will tell you the achievable cancellation ratio in the sidelobe cancellation system, with aperture A, bandwidth B, from a jamming direction of θ. Coincidently, when θ is equal to 0, that is the jamming is normal to the array, then there's no time delay across the aperture and, the sidelobe cancellor is essentially perfect.

Now the premise here was that if you had N degrees of freedom, one hypothesized N jammers. If you have more degrees of freedom than you do jammers, then things get better, which I'll describe in a moment. Slide 9 In fact, when we talk about conceptual ways to cope with the broadband problem, Slide 9 illustrates the candidate ways that people would talk about at present. One is to use additional degrees of freedom. If I had twice as many auxiliary channels as I do jammers, then things would be much better. Or one could put tap delays comparable to some reasonable fraction of the free space aperture, and increase the dimension of only the adaptation process. In effect, I have corrected the time delays across the system.

I will be talking mostly about sub-banding as a technique. By sub-banding, I mean you have a very wide bandwidth and you break that bandwidth into a bunch of sub-bands and then you independently adapt in each one of those sub-bands which are now narrowband and then you combine the result of the output and generate the adapted wide bandwidth output. Finally, exploiting the very rapid convergence properties that I've talked

about before, for some waveform applications like spread spectrum and specifically frequency hopping, waveform design is the most powerful technique of all, in as that one can literally adapt narrowband. I'll discuss more about that later.

The first technique that we talked about was additional degrees of freedom. To understand that, I will elect to use the artifice that I talked about before Slide 10, i.e., a point source in the broadband system was mathematically equivalent to narrowband system where the jamming is spatially spread as shown in Slide 10. So the jammer is effectively spatially spread over the shape of whatever the receiver pass band is in relative amplitude. Now suppose we have an aperture, and two elements, and only one jammer. If you forget about the aperture and just ask what pattern potentially could be generated by those two elements, it would be some sort of interferometer pattern. And the relative phase between those two could adjust the null to any position you want. If you take that pattern from these two elements, adjust it to where the zero is you want, and apply weight to the amplitude and phase, you can further adjust the slope. You can in a sense, match both one point and the slope of the pattern, and therefore you can do much better over a wider bandwidth.

Most systems will have time bandwidth products less than one half, because if you approach a time bandwidth product of 1 that really means that the full time of the antenna is such that you can't coherently combine the energy anyway. This is because energy coming into one sideband antenna isn't there when it combines with energy from the other side of the antenna. Typically for time bandwidth products less than one half, two auxiliary channels per jammer seem to be adequate. So if you have 5 adaptive degrees of freedom for a narrowband system with 5 jammers, then you're required to have 10 auxiliaries with that approach.

Another approach is to use time tap delays. This has been used primarily in systems that have been retro-fitted where something probably didn't work, because it is difficult to add new adaptive antennas and receivers to a system but not as difficult to add taps to a system. Slide 11 The first attempt to do this is to apply the type of tap where the main beam had a tap position at a somewhat arbitrary delay of T/2 and then you apply 2 adaptive degrees of freedom per channel on the auxiliaries, one on each end See System A Slide 11. That works fine for jammers with time bandwidth products of less than one half that are off boresight. But if the jammer happens to be normal with the array, then again there's no solution. So practically, one needs at least two taps for every channel including the main beam for this technique to be viable See System B Slide 11. Furthermore simulations show that whereas the technique works, the achievable cancellation ratio is dependent upon the geometry of the jammers themselves.

Slide 12 The technique that I'll be talking about primarily are sub-bands. Now the disadvantages of this technique are shown in Slide 12. You have to sample at a somewhat higher rate than the bandwidth of the system. But this would be true anyway if you're trying to achieve very high levels of cancellation. At a certain level, the receiver matching dominated the achievable cancellation ratio, and many digital systems require very careful matching between the inphase and quadrature components of the digitally implemented complex output, as that class of errors is non-linear and cannot be corrected by the adaptive system.

However, what it does do is this

Since you are basically sampling very near the instantaneous bandwidth of the system, it does not require any additional antennas or receivers. This is because you are just taking samples in the normal bandwidth of the system and you are just operating on them in the digital domain. Furthermore, the computational rate that is required to do sub-banding, ends up being exactly what would be required in the narrowband system. There's additional memory but it is not more computationally burdensome than would be a narrowband system. Since for many cases, it turns out that receiver matching is often the limiting factor for achieving high levels of sidelobe cancellation, this sub-banding technique will automatically align the receivers for you.

Slide 13. This is in effect what we are doing. See Slide 13. Suppose we have two channels, channel 1 and channel 2, the receiver pass bands are not exactly the same in the wideband system. When you try and cancel one from the other, you can only cancel 1 point perfectly, and there'll be a residue that's a function of how well the receivers are physically matched. Now let's suppose we are sub-banding the channels by a factor of 8. When we sub-band, what we are conceptually doing is generating 8 filters across the pass band of the 2 receivers, so we'll be independently adapting in the same filter over the number of degrees of freedom that we have. Now we have the capability of perfectly correcting an amplitude and phase, except for any little distortion over a much smaller portion of the band, given that you have suitably designed filters with sufficiently low frequency sidelobes.

A drawback, however, is that real world receivers are not purely bandlimited. The receiver response of Channel 2 on Slide 13 is the square wave of the frequency domain and it has tails. An example I have in a sampled system is that filter F1 will repeat at some alias position that went over the sampling rate. Unless this alias occurs sufficiently far down on the receiver response, when you try to adapt, you cannot. You can only generate one adaptive weight and therefore you cannot do perfectly on both sides of the receiver characteristic. So what that means is that one must oversample sufficiently to assure that the aliasing by making $T=1/\delta F$ smaller so that the aliasing position is way down on the receiver response. With that condition this technique works very well.

Conceptually Slide 14 illustrates how the block diagram would look utilizing that approach. We have an adaptive beam with N degrees of freedom. The E_i's would be antennas going through receivers and A-D converters. Suppose we are sub-banding by a factor of R. What one would do is to take R contiguous samples, form an FFT on each of those channels, and filter by filter, one would invert an N+1 by N+1 covariance matrix and solve the problem for each of the sub-bands. Then you combine the information of the N by taking the inverse FFT.

I said that computationally this is not burdensome, and one way to look at it is to visualize when we adapted wideband and did not sub-band. Or it could have been a narrowband system where the bandwidth is $1/\tau$ and τ is some pulsewidth. The algorithm we alluded to earlier was that 2N samples are adequate in terms of time to get a nearly optimum solution. So the time base of this process is $2N\tau$ duration which is a very small period of time. If a certain number of operations have to be performed in that period of time, then one could ask, "Well, how many computations do you have to do per unit time?" The time would be $2N\tau$. Yet there would be the number of computations divided by this time and that would be the computations of the

instantaneous bandwidth of the system.

Now if you would use R contiguous samples to sub-band, that's as if you transmitted a pulse that was $R\tau$, making the bandwidth $1/R\tau$. Now 2N samples is 10R times as long, and even though you have to do R-independent matrix inversions, the computations per unit time are exactly the same because the computations of the original bandwidth are exactly the same. This is because the time base is also extended by a factor of R.

There are architectures that could actually make the adaptive process completely ignorant of the fact that you are doing this sub-banding at all. That is, once you have an architecture that can solve the problem wideband, then you make the system blind to that, i.e. you can have FFTs or you can time-multiplex the inputs, so that the output will be adaptive and the adaptive array processor will be blind to that. It will think it is looking wideband with adaptation in effect.

Slide 15 is an example where you are sub-banding by a factor of 4. The channel samples number in batches of 4 and puts them in the processor. It performs FFTs. Now everything with a double square around it in Slide 15 is hardware that you have to add to the system as the penalty for sub-banding. You will have to have an FFT on each channel at the input and inverse FFT at the output. Computationally, whatever mechanism you use for inverting or implicitly solving these systems of linear equations is exactly the same except for some multiplexing circuitry. However, the memory does increase by a factor of R, or as a sub-band in this case, a factor of 4.

A fourth technique that I talked about was frequency hopping. Suppose you have the system which gets its broadband by frequency hopping, i.e. put out pulses at a given frequency and then some other frequency, etc. Just for illustrative purposes, let's talk about an example of the step version of linear chirp, that is where the delta frequencies are contiguous in frequency as shown in Slide 16. In principle they wouldn't have to be. The number 2 pulse could be at some frequency that's much further away than the reciprocal bandwidth. To generalize it, we could talk about a coded pulse at the given frequency such that the bandwidth is $1/\tau$ at some pulse compression output as the basic bandwidth of the system. At some later period of time we are going to output another frequency. These could be contiguous in time in principle. Now analog loops could never exploit this because what you're doing in an analog or iterative system is applying the adaptive weights that you drive to new data and the convergence is very slow. Therefore, if you adapted one frequency it is not applicable to some other frequency. You can't presumably assume another frequency. What you can do, from what we have simulated, is the technique of applying the same data to the covariance matrix from which the data was formed. What that means is that as long as one does not hop to a new frequency, i.e., if the time between frequency hops divided by the pulse width, or really that time bandwidth product, is greater than 2N, where N is the number of adaptive degrees of freedom, then you add adequate samples of the given frequency to solve the problem. If you store the data then there are no holes in the data. So that means in this scheme you can literally adapt narrowband. The bandwidth can be orders of magnitude less than the total bandwidth of the system in this case. If the frequencies are contiguous with total control, then there'd be reduction by a factor of S where S is the number of frequency hops.

To summarize what all this means

computationally, in terms of hardware, see Table I, Slide 17 you go back to a narrowband system where we do not have a bandwidth aperture or time bandwidth product problem. In order to cope with any jammers within the antenna, the receivers and A-D converter, we will hypothesize that we use a sample covariance matrix approach. There are many equivalent algorithms to do that of course, and we require on the order of $2N^3$ computations. But that would subtend over 2N samples of the bandwidths of the computational rate of instantaneous bandwidth. If we divide BW by 2N, this would be the number of computations at the instantaneous bandwidth of the system. The amount of memory is $N^2/2$ that you need to physically store the data associated with this processing. These are complex words - inphase and quadrature. If one elects to reprocess the same data then the memory increases.

We talked about four techniques with broadband system. If we use extra auxiliaries, you will need somewhere between 2 and 3 times as many auxiliaries as you have degrees of freedom depending upon the time bandwidth products With this the real penalty is hardware The antenna receivers A D s are analog and the computations and memory are digital. One usually argues well you can always build a number cruncher, but you don't want to increase the analog circuitry hardware in the system. What this does is increases the total computational loading in this example by a factor of 8 and the rate by a factor of 4. If the time bandwidth product is less than a half this will work with a fairly wide set of conditions.

If you use tap delays for time bandwidth products of approximately one half, you need on the order of three taps per channel. You only need 1 antenna and 1 receiver of course. However, now you need 3N A-D because normally the tap delays will correspond to the actual free space fraction of the free space distance across the array The instantaneous bandwidth in the A-D converter will be less than that, and therefore you will need to have faster A-D converters, or often more A-D converters. If you have 3 tap delays, the covariance matrix you're inverting is a 3Nx3N array instead of an NxN array.

If you sub-band by some factor of R, the analog circuitry does not increase at all. The total computations increase but the computations per unit time are the same. However the memory does indeed increase by the sub-banding factor.

Finally, for the specialized case of the frequency hop waveform, things are orders of magnitude better because now literally the computational rates are reduced on the order of the square of the number of frequency hops over what would have been in the broadband system

I want to show how this relates to the receiver equalization problem If you have receivers that are mismatched, and you subtract one from the other you have receiver residue. We have worked already on an HF array and what we did is we built equalization network array where conceptually you could put tap delays You could use one receiver as a reference and put tap delays behind the others and form adaptive weights. So what you are trying to do is make every receiver match every other receiver. In doing that we were able to get sidelobe cancellation in the system. I think it is the most anyone has ever achieved; it's 63 dB in that system utilizing 16 tap delays, about twice as many taps as the number of poles, using Butterworth filters. The point is that you don't have to do that when you sub-band. The reason you don't have to do it with sub-banding is because we now have multiple narrowband filters across the band

Therefore when you are adapting the sub-bands you can tolerate much more receiver mismatch since you could now match its amplitude and phase in different parts of the band that won't interfere with sampling in other parts of the band. So there is the synergistic effect, and receiver matching is not the dominant source of sidelobe cancellation. The wave of the future is not 20 to 25 dB. the correct state of the hardware applying digital techniques in the field will support 40 to 50 dB for sidelobe cancellation And this will probably be in systems in the near future.

SIDELOBE CANCELLOR TECHNIQUES

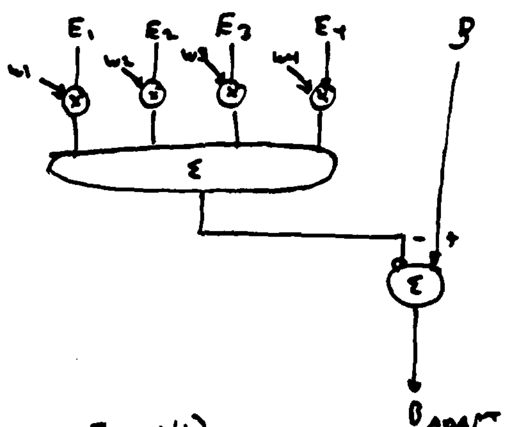
IN WIDEBAND SYSTEMS

SLIDE 1

ADAPTIVE SENSORS, INC.
216 PICO BLVD., SUITE 8
SANTA MONICA, CA 90405

FIGURE 2

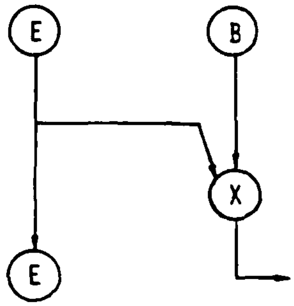
SIDELOBE CANCELLOR



$$B_{ADAPT} = \bar{E}^T M_E^{-1} \begin{pmatrix} 1 \\ 0 \\ 0 \\ 0 \end{pmatrix}$$

SLIDE 2

(U) SINGLE CHANNEL SIDELOBE CANCELLOR



$$B_A = B + wE$$

GRAM SCHMIDT ORTHOGONALIZATION

$$w = -\frac{B\hat{E}^*}{|\hat{E}|^2}$$

$$\hat{B}_A E^* = 0$$

SLIDE 3

ADAPTIVE ARRAY SPEED LIMITATIONS

ANALOG	TYPE		SPEED
	• APPLEBAUM HOWELL	MAX S/N	• EIGENVALUE DEPENDENT
	• WIDROW	LMS	• NOISE FIELD DEPENDENT

DIGITAL

- SAMPLE COVARIANCE MATRIX

$$\hat{M} = \hat{E}E^*$$

$$\hat{M}^{-1}\bar{S} = \bar{W}$$

$$B^*_{ADAPT} = E^*\bar{W}$$

M = STEADY STATE COVARIANCE MATRIX

$\hat{M}$ = SAMPLE COVARIANCE MATRIX

K = # SAMPLE

N = # ADAPTIVE DEGREES OF FREEDOM

- COVERGENCE PROPERTIES

$$R = \left[(S/R)_K \text{ SAMPLES} \Big/ (S/R)_\infty \text{ SAMPLES} \right] = \begin{cases} \frac{K-N+1}{K+1} & \text{NEW DATA} \\ \frac{K-N}{K+1} & \text{RE-PROCESSED DATA} \end{cases}$$

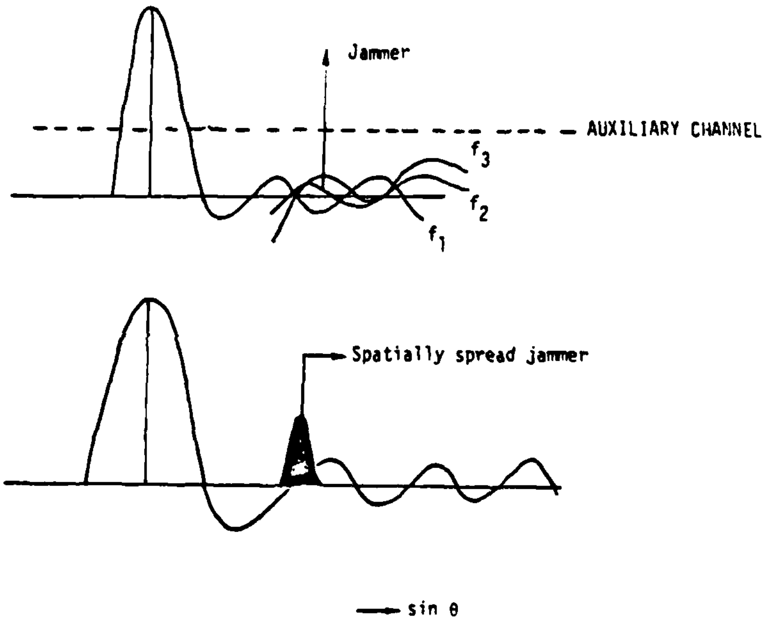
- R IS INDEPENDENT OF M
- R WITHIN 3dB AFTER (2N) SAMPLES

SLIDE 4

BANDWIDTH-SPATIAL SPREADING EQUIVALENCE

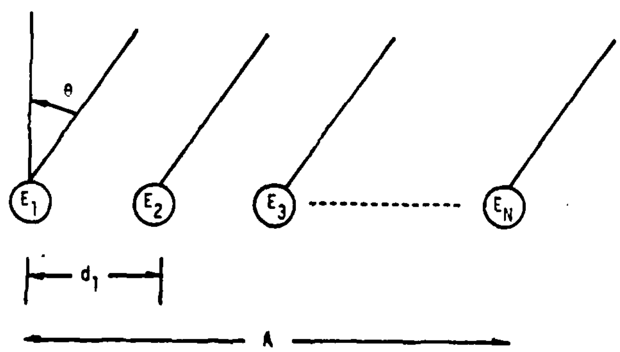
BROADBAND POINT SOURCE ⟺ NARROWBAND SPREAD SOURCE

(A) (B)

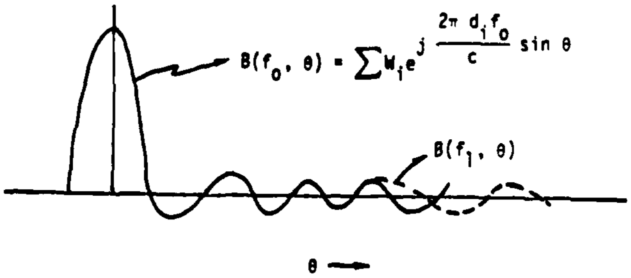


SLIDE 6

WIDE BANDWIDTH PROBLEM



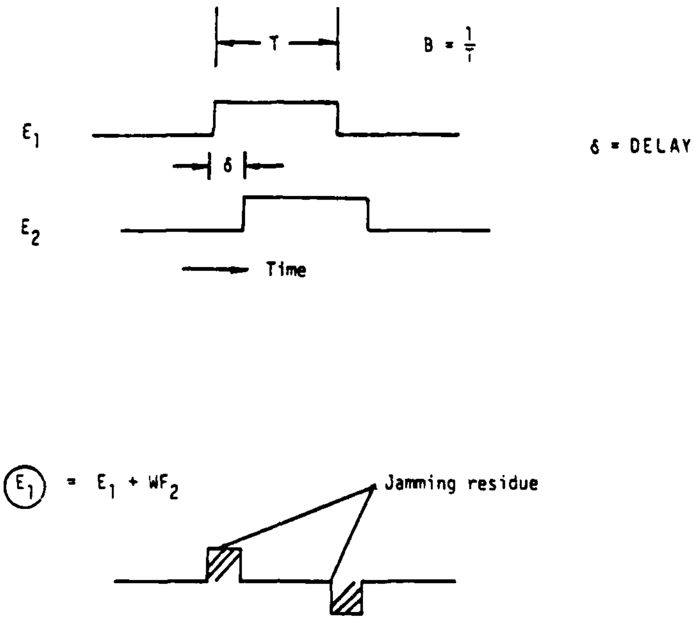
$$B(f_o, \theta) = \sum W_i e^{j\frac{2\pi d_i f_o}{c}\sin\theta} \qquad \text{Eq. 1}$$



$$B(f, \theta) = B(f \sin \theta)$$

SLIDE 5

TIME DOMAIN SLC LIMITATION



SLIDE 7

WIDEBAND INTERPRETATION SUMMARY

EQUIVALENT SPATIAL SPREAD $\qquad \frac{d\theta}{df} = \frac{\text{Tan }\theta}{f}$ $\qquad$ Eq. 1

Absolute $\qquad \Delta\theta = \frac{B \text{ Tan }\theta}{f}$ $\qquad$ Eq. 2

Normalize (Fractional Beamwidths) $\qquad \frac{\Delta\theta}{\theta} = \frac{BA}{c}\sin\theta$ $\qquad$ Eq. 3

TIME BANDWIDTH PRODUCT

$$P = TB = \frac{A\sin\theta}{c}B \qquad \text{Eq. 4}$$

SIDELOBE CANCELLOR LIMITATION

$$CR = \frac{1}{12}\left(\frac{\pi A\sin\theta B}{c}\right)^2 = \frac{\pi^2}{12}P^2 \qquad \text{Eq. 5}$$

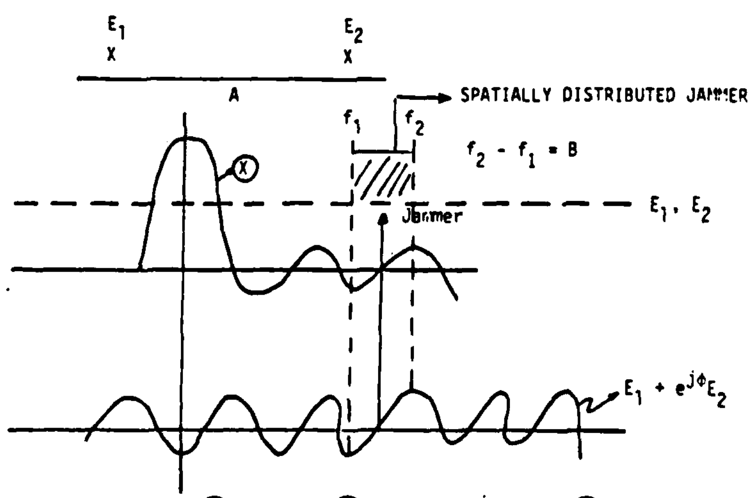
SLIDE 8

BROADBAND ADAPTATION TECHNIQUES

- Additional Degrees of Freedom
- Tap Delays
- Sub-Banding
- Waveform Design

SLIDE 9

SLC SOLUTION WITH ADDITIONAL DEGREES OF FREEDOM



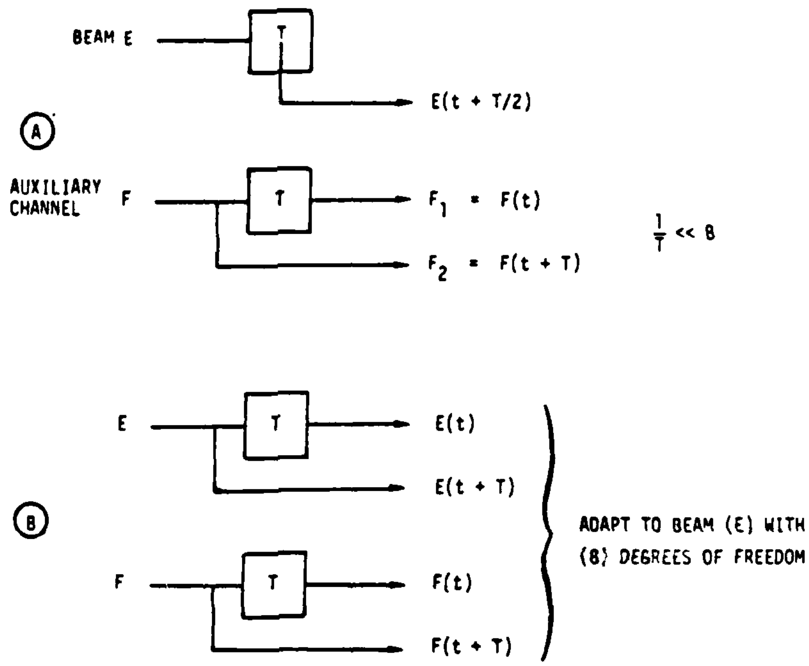
$$\textcircled{X}_{ADAPT} = \textcircled{X} + w(E_1 + e^{j\phi}E_2) = \textcircled{X} + w_1E_1 + w_2E_2$$

$$P = BT = \frac{BA \sin\theta}{c}$$

For $BT << 1$	# AUX = Jammers	$SLC = \frac{\pi^2}{12}\rho^2$
For $BT \leq 1/2$	# AUX = 2 x (# Jammers)	

SLIDE 10

TAP DELAY IMPLEMENTATION



SLIDE 11

SUB-BANDING TECHNIQUE

DISADVANTAGES -- • REQUIRES SAMPLING > B

• REQUIRES I, Q MATCHING

ADVANTAGES -- • AVOIDS ADDITIONAL ANTENNAS, RECEIVERS, A/D'S

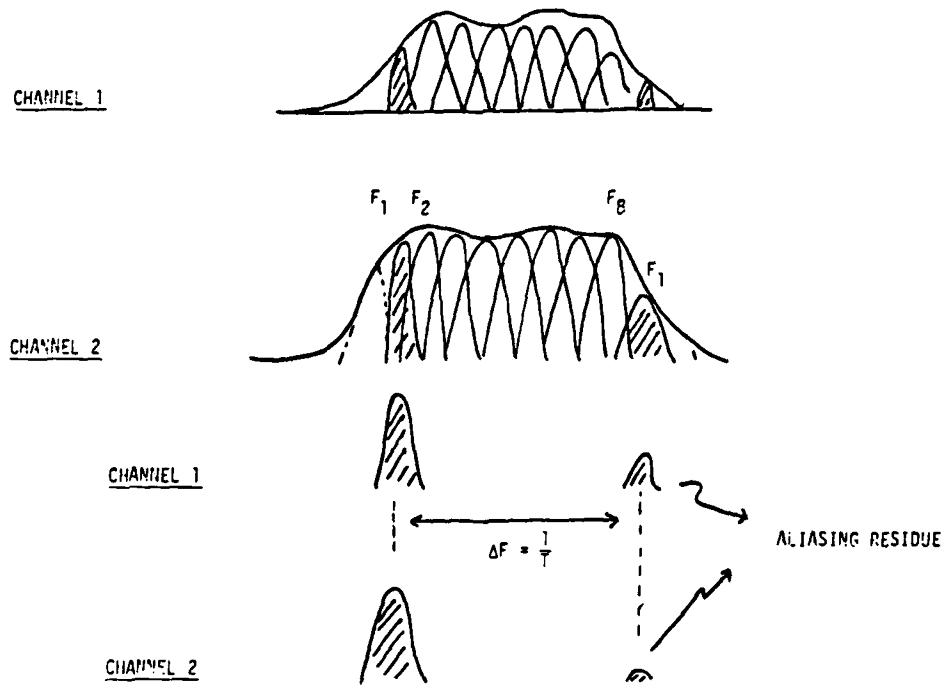
• COMPUTATIONAL RATES INDEPENDENT OF SUB-BANDING BANDWIDTH

• SYNERGISTIC RECEIVER MATCHING

• REDUCES EFFECTS OF MULTIPATH

SLIDE 12

SLC SOLUTION WITH ADDITIONAL DEGREES OF FREEDOM



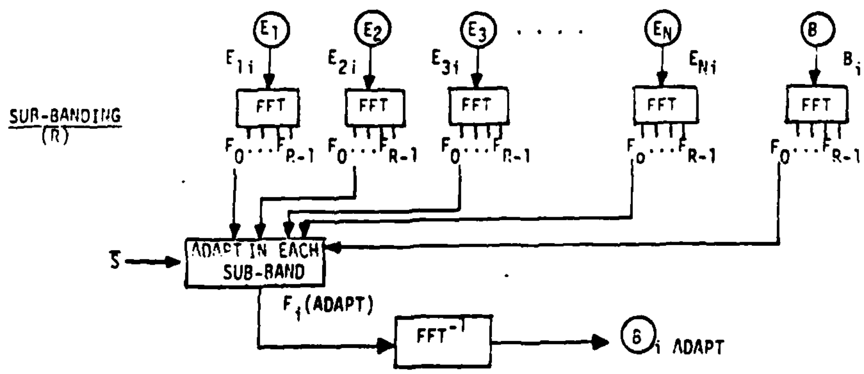
(1) MATCH RECEIVERS

OR

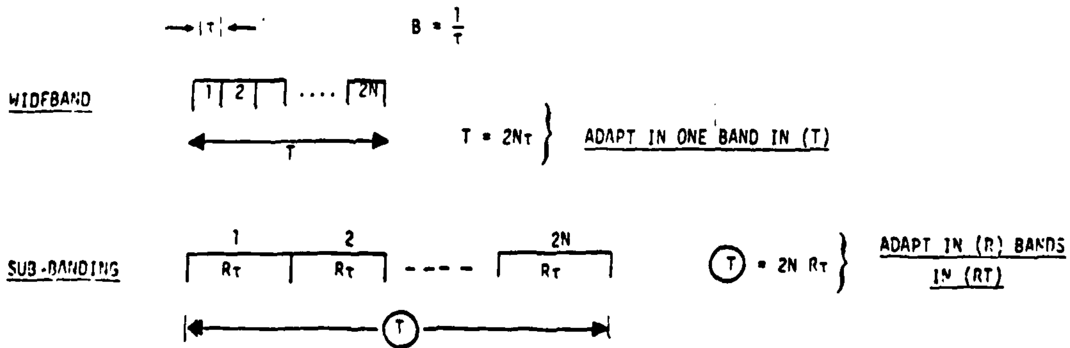
(2) OVER-SAMPLE

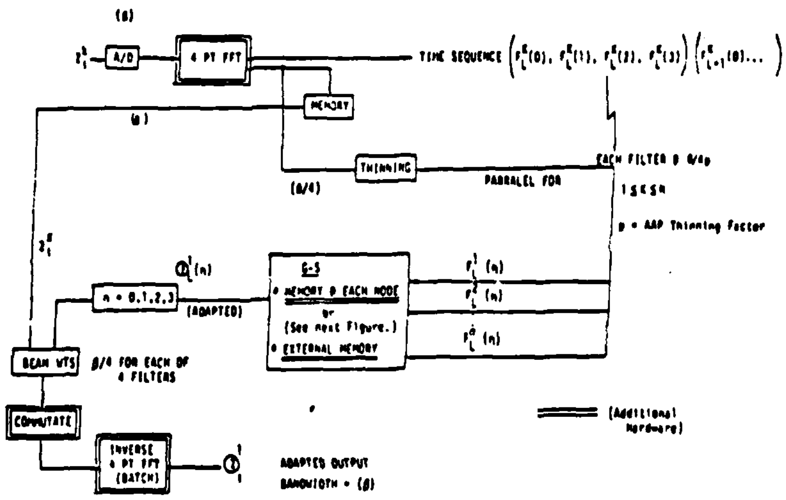
SLIDE 13

SUB-BANDING TECHNIQUE SLIDE 14



COMPUTATIONAL INVARIANCE



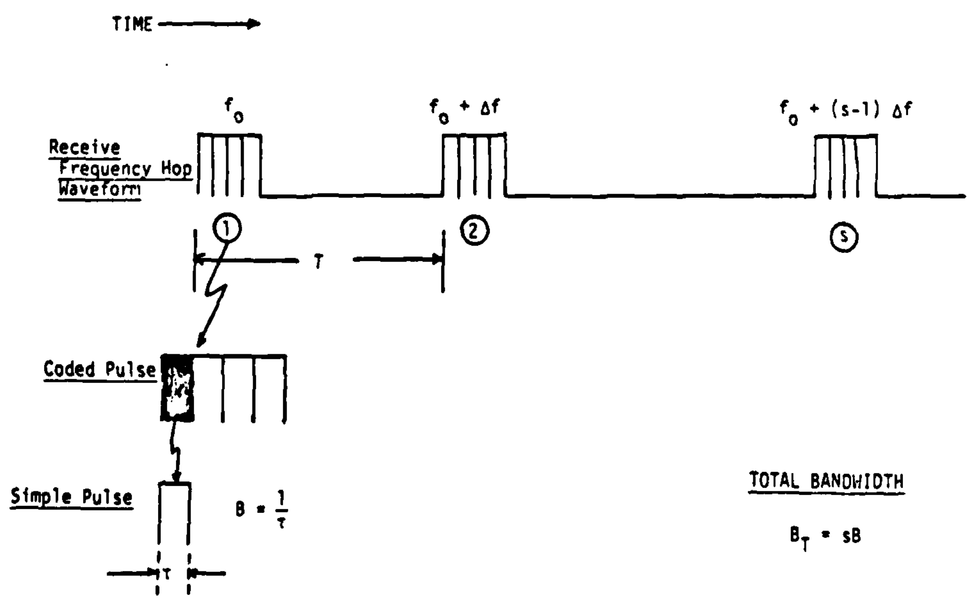


Bandwidth Reduction -- AAP Architecture

SLIDE 15

FREQUENCY HOP WAVEFORM

(Example -- Step Frequency)



TOTAL BANDWIDTH

$B_T = sB$

Radar $R_u \leq \frac{CT}{2}$

R_u = Range uncertainty

Communications $R_u \leq CT$

SLIDE 16

TABLE I

HARDWARE COMPARISON SUMMARY

(N Jammers)

	Antenna	Receivers	A/D's	COMPUTATIONS Batch	COMPUTATIONS BW	MEMORY New	MEMORY Reprocess	Comments
NARROWBAND	N	N	N	$2N^3$	N^2	$N^2/2$	$\frac{5}{2} N^2$	Narrowband Reference
BROADBAND								
• xtra AUX	2N 3N	2N 3N	2N 3N	$16N^3$ $54N^3$	$4N^2$ $9N^2$	$2N^2$ $\frac{9}{2} N^2$	$10N^2$ $\frac{45}{2} N^2$	(TB < 1/2 Wide set of Conditions)
TAP DELAYS (3)	N	N	3N	$54N^3$	$9N^2$	$\frac{9}{2} N^2$	$\frac{45}{2} N^2$	(Geometry Dependent)
(SUB-BANDING)(R) (R Sub-bands)	N	N	N	$2RN^3$	N^2	$\frac{RN^2}{2}$	$\frac{5}{2} RN^2$	(Synergistic Effects)

SLIDE 17

JAMES DUPREE

Slide 1 I want to talk about multiple access adaptive antennas and give two examples to motivate this discussion. Primarily my talk will be about the TDRSS application, since I've been living with that for the last 8 years, and we have now almost got to the point of having an opportunity to test it. (Laughter.......)

Slide 2 TDRSS uses an S-band array of 30 helical elements on each of 3 relay satellites with beamforming on the ground. We use spread spectrum multiple access techniques to accommodate up to 20 user satellites. These beams for the user satellites are independently steerable over 26 degrees field of view and they provide 30 dB gain for each of the users at S-band. We use an adaptive maximum SNR algorithm for calibration. This is different than the max SNR algorithm that you've heard discussed here today. I'll go into that. We also use orbital models to do open loop beam steering between the adaptive calibrations.

The other example to consider is one in which multiple access is provided by an area coverage beam. That is, we have to provide a beam steering vector to produce a quiescent pattern which a number of users can randomly share on either a frequency multiplex or time shared basis. Normally, you would hope to have low sidelobes outside the area of coverage and use frequency hopping spread spectrum techniques. Economics of the satellite application drive you toward the millimeter wave applications, and one would normally use a power inversion algorithm for high resolution jammer nulling.

Slide 3 The basics of TDRSS are a system of tracking and data relay satellites in geo-stationary orbit, together with the ground terminal at White Sands, NM, whose purpose it is to replace the worldwide ground stations Figure 1 shows the network. Slide 4 There is a forward link established from the ground station to user satellites and the user satellites respond with a return link to the ground station. There are a number of different services provided in the system The particular one that we want to talk about is the multiple access S-band service to low earth orbit user satellites through the relay and down to the ground. Slide 5 For this purpose, the satellite carries this 30 element helical array on the body of the satellite. See Figure 2. In addition, there are other antennas. There is a dichroic-feed shared dish which contains S-band and K-band single access services (Again, see Figure 2.) We'll concentrate on the multiple access application. Also, the ground station at White Sands has 3-60 ft. K-band dishes to track each of the 3 satellites, the East and West satellites and the Central Satellite See Figure 3.

Slide 6 The network diagram of operation looks like this see Figure 4. The user satellites transmit at S-band to the relay satellite which then FDM multiplexes the element signals to the ground via K-band. On the ground, there's a 30 element FDM demux which retains the phase information on this downlink by means of the pilot tone. Since there is a beam formed for each of the user satellites, there are 20 30-element beamformers. We go out of the beamformer into an acquisition and demodulation system which includes the PN despreading. In this system we must do our PN acquisition despreading following the beamformer weighting. However the adaptive controller has to use the spectrum spreading information on a beacon signal in order to determine and identify that signal and adapt to it. This is because in this application we do not apriori know exactly the direction of arrival and/or phase information entirely of the system. The primary purpose of the MA-calibrator is to

estimate the beam steering vector.

Slide 8 The PN spreading sequence chip rate is 3 megabits/second, and we are able to accommodate data rates of 100 bits to 50 kilobits per second at S-band in a .26% bandwidth. The purpose of the PN spreading is, in our application, to satisfy the CCIR guidelines by reducing the flux density incident on the earth. Besides that, the PN sequence provides user identification, reduces the mutual interference, rejects narrowband interference and provides accurate ranging.

Slide 9 Remote beamforming works like this. The 30 element channels are coherently telemetered to the ground station, where the 20 beamformers each combine the 30 element channels to track the users. Therefore, the control complexity is in the ground, not in the spacecraft. We use a pilot tone technique for downlink phase referencing, but we have to cope, however, with ionspheric time delay variations since we do not estimate those. That is where the adaptive system comes into account. Also, random phase and gain drifts occur in the satellite and therefore, we need some type of calibration. A simplified diagram (in Figure 5) shows the FDM multipler Slide 10 taking the 30 element signals through the return processor, translating to K-band and downlinking to a receiver where there is a cascade of demultiplexors which undo these operations and go into the MA beamformer. The MA demodulation, bit sync and error correction decoding occur following the beamforming.

Slide 11 The problem is how to calibrate and control the beamformers, track moving user satellites in low earth orbit, and we have to do so, contractually, within 8/10 of a dB of the theoretical SNR achievable by an adaptive array. In this case, with thermal noise being the primary consideration, the output SNR has to be the sum of the SNRs in each of the elements.

Slide 12 Our control options were: open loop control, which won't work because of the ionosphere; full time adaptive control which would require 20 adaptive loops and is very expensive; time-shared adaptive control which is not fast enough; or open loop prediction with time shared adaptive calibration which is what we opted for. I'll try to explain that concept shortly. Slide 13 We do open loop pointing with a main frame computer which estimates the optimal weight vectors from known satellite ephemerides, the orbital model, and calibration vectors which are supplied to it. A closed loop adaptive control system periodically updates the weight vector for a known source location, with the calibration being used to reset the initial conditions for the open loop pointing program.

The generalized concept for an adaptive control system, as we have implemented it, is like this see Figure 6. A user signal impinging on the array is subjected to a vector transfer function, Xi. An interfering signal, which we assume could exist in our system, is similarly transformed by vector transformation, Xi. There, we have received vectors, $\underline{r}$, which include thermal noise, as inputs to the beamformers. The output of the beamformer is a scalar which we could represent as an inner product, $(\underline{w},\underline{r})$. Each of the input element channels is sampled and then correlated with the inner product scalar output to produce correlation vectors. These are updated by a gradient algorithm which produces the weight vector $\underline{w}$. My mathematical convention is to conjugate the weights in the beamformer in order to counteract the phase slant caused by the angle of incidence. Similarly the inputs to the correlators are conjugated. Slide 15 We have seen Eq. 1 before as the max SNR. (See Eq. 1 on Slide 15.) You could prove

by the Schwartz inequality that in fact the weight vector in Eq. 2 satisfies the conditions for max SNR. The open loop *prediction technique* works as follows. Once I have a beam steering vector, and I know the system dynamics, I can generate a transformation matrix that will produce another beam steering vector and thereby, step the beam. The adaptive *calibration* occurs because in the steady state, the loop must approximately satisfy Eq. 2, and this turns out to be a correlation vector which I can measure. Therefore, I can estimate the beam steering vector and that vector will not contain nulls in it. That is a handy thing for open loop predictions because you're not really able to handle the nulls in the open loop prediction.

Slide 16 seems like a vast jump from the slides that I had previously, but I must explain that since my problem was to generate a gradient algorithm, I needed to know the gradient. Since I like to work in complex coordinates, I had to also cope with the fact that the SNR is a real scalar function of a complex weight, and I run into analytical problems if I try to apply the standard types of definitions. Therefore, I've worked out a method in which I use the dual identity of complex numbers as either elements of a complex vector space or as members of the real vector space of ordered pairs of real vectors. I must do that in order to be able to have a satisfactory definition of derivative or differential, because I have to have linearity and continuity with respect to an increment h. In this application, my criterion is SNR. Also I want to apply an increment to the weight vector at some given point and see how much the SNR has changed and from that, estimate a term which is linear and continuous in the increment h such that the continuity relation in Eq. 1 on Slide 16 is satisfied. So I take the transformation $T\omega$ as the ratio

$$T\omega = \frac{(\omega,A\omega)}{(\omega,B\omega)}.$$

Slide 17 I do a little trick on that by jumping back and forth between the identities of the vector space, V^{*}, which could either be considered as a real vector space of pairs of real vectors, or as a complex vector space. Of course I have to also make a similar kind of jump in the way I *derive* the matrix A, but I have found that this relationship does hold. There is a linear function which turns out to be:

$$\delta T(\omega;h) = 2RE\frac{(A\omega-TB\omega,h)}{(\omega,B\omega)}$$

Therefore, by a little Schwartz inequality argument, one can show that the complex gradient which maximizes the norm of the Frechet differential is in the form $g=A\omega-TB\omega$. Thus stationary points of the SNR occur at eigenvectors of the eigenvalue problem $A\omega=\lambda B\omega$ which for convenience, A could be considered to be a signal covariance matrix and B could be a signal plus noise covariance matrix.

Slide 18 So the max signal to noise ratio algorithm for the eigenvalue problem appears in the form $\omega_{n-1}=\omega_n+\mu_1(\ell_n A\omega_n-B\omega_n)$. This is very similar to the eigenvalue problem residual except that I've factored out the λ and put it in the step size u_l. The result is, in effect, decreasing the step size. As the SNR improves, the step size gets smaller and I worry less and less about the effects of fluctuation noise and upsetting the measurements that I have made on the weight vectors. In addition to that, ℓ_n is the ratio of some constant which could be the output power reference divided by signal power,i.e.

$$\ell_n = \frac{C_n}{(\omega_n, A\omega_n)}$$

So this is an AGC amplifier and then there is a normalization equation which determines what the output power reference should be set at in order to obtain normalization of the weights, i.e. $c_{n+1} = C_n + \mu_2(R - \|\omega_n\|^2)$. These 3 factors are important in determining the convergence of the algorithm.

Slide 19 Figure 7 shows the implementation. We see the received array vector which comes into the beam former and appears at the output to be despread by a PN code on a local oscillator. The PN code is derived by an acquisition procedure of the system at the output of the beam former. There's a bandpass filter which in our case has a digital implementation. The bandpass filter is also used in the signal path and the sum path output goes through the AGC amplifier. The conjugation and complex correlator occurs after the BPF and that provides the signal component of the SNR gradient estimate. There's a band reject filter, or in our actual case, there is a wideband bandpass filter which passes both noise and signal through another correlator with negative feedback for that loop. There's a simple update algorithm in the computer. Not shown in the simplified diagram are the loops for weight normalization feeding back into the AGC amplifier for control. That's basically the implementation.

In our system, to point out how the 20 beam formers are done, see Figures 8 and 9. After the FM demux there is a 20 way power divider going into 20 parallel beam formers and the adaptive loop must have the ability to be connected to any of the beam formers. It turns out to be a horrendous job of cabling and switching. We also have the problem that any of the 3 satellites must be able to be connected to any one of the 20 beam formers, and the adaptive loop must be able to connect up to any of them.

Slide 20 The control and calibration concept that I've outlined here, could be generalized, as shown in Figure 10. The idea is that I have a closed loop adaptive control model which might in the general case also incorporate system dynamics. This could be a very low bandwidth loop which would be able to nevertheless keep up with the satellite because the user dynamics would be known apriori. This is the idea of incorporating system dynamics, a la the Kalman filter. There is a calibration update model in which apriori estimates of the beam steering vector are compared with correlation vector estimates to derive, via Kalman filters, an update of the open loop weight estimation model. In the absence of the adaptive calibration, this loop continues to run and propagate the beam steering vectors using the system dynamics. Slide 21 The conclusions are: we use the system dynamic model to reduce control loop bandwidth, we reduce complexity by time sharing the adaptive control loop and, I believe this has applications to avionics and laser beam steering as well as satellites.

Slide 22 In the beginning I had touched on an example in which we have a tactical beam steering vector estimate. Its fairly obvious that instead of estimating the beam steering vector, one might stick it in apriori and in that case, one would have to only adapt to the noise sources that appeared in the external world. A problem in that application is when using an MBA one might say, use all ones as the beam steering vector as opposed to phased array applications which used a one and all zeros for the nulling elements. That doesn't answer the question precisely because as some researchers at MIT have shown, there are advantages to introducing phase information into the beam steering vector. It turns out to have

a dramatic impact on the resolution one obtains by nulling, particularly in using the MBA. I ask the question of what should the criterion be for beam steering vector phase? It should not be ad hoc. If one has multiple users in the area, perhaps the users do not have all the same information rate and therefore do not need the same gain. How does one go about optimally allocating gain for multiple user community? What is the function of the antenna, how does one describe the antenna as an information channel? Perhaps one uses multiple access information theory.

In the area of performance prediction, I find that, if I take into account things like weight tolerances, I want to know what is an effective choice of bits for the digital beam former. I don't really have a way of analyzing this except for the case in which the signal and noise covariance matrices commute. In that case the SNR distribution is a singly non-central F distribution. This is opposed to the case in which I would be trying to estimate the sample covariance matrix in the direct matrix inversion technique, in which case the SNR distribution is known. This is a slightly different case. Also, convergence speed analysis can be done for some simple cases. e.g. when we have thermal noise covariance. It cannot be done in the general case, I think, in which the matrices do not commute

Digital implementation and satellite application puts processing throughput on a critical path. We need an understanding of the things that have been said about sample matrix inversion techniques and digital processing. We must understand that one has to argue with program managers that the system that we're going to build will weigh, including the antenna, less than 100 lbs. and will consume 10's of watts. Those are real life facts. It is true that array or network computer architectures based on matrix factorization might yield some dramatic improvements in convergence time. However, we are not terribly hurt on convergence time at the moment unless one gets into deep consideration of the optimum strategies against these threats: pulse jammers, partial band jammers or frequency following jammers.

Slide 23 G/T needs cry out for low noise millimeter wave amplifiers over very large bandwidth. You can set the G/T performance of your antenna. Without that, you are forced into doing RF beam forming. 16 bit computers with 1-2 megahops throughput (fixed point), low power (10-20 watts) and radiation resistance on the order of 1 million rads, total dose, would be very helpful. I think this is almost a must now for satellite applications. We are going to forward error correction memories. We discussed digital beamformers. I think that's possibly applicable to TDRSS, e.g. if we had the inner product chip that could do the digital beam forming, it would be great. Possibly, in the far future, the parallel processing of optical adaptive algorithm would replace all the gyrations that we are going through to do digital processing at the moment.

I have a slide on an approach to the proof of convergence of the max SNR algorithm, and also some written material on the discussion of the beam former or the beam steering vector which are supplied.

MULTIPLE ACCESS ADAPTIVE ANTENNAS FOR COMMUNICATION SATELLITES

JAMES E. DuPREE

SLIDE 1

MULTIPLE ACCESS ADAPTIVE ANTENNA EXAMPLES

TDRSS--

- o S-BAND ARRAY OF 30 HELICAL ELEMENTS ON EACH OF 3 RELAY SATELLITES.
- o GROUND-BASED REMOTE BEAMFORMING.
- o SPREAD-SPECTRUM MULTIPLE ACCESS TECHNIQUE, FOR LOW FLUX DENSITY.
- o UP TO 20 USER SATELLITE BEAMS INDEPENDENTLY STEERABLE OVER 26° FOV.
- o 30 DB GAIN PER BEAM, AT S-BAND.
- o ADAPTIVE MAXIMUM SNR ALGORITHM FOR CALIBRATION.
- o OPEN-LOOP BEAM STEERING, USING ORBITAL MODELS, BETWEEN CALIBRATIONS.

MILITARY TACTICAL COMMUNICATION SATELLITE--

- o TACTICAL AREA COVERAGE BEAM WITH FLAT GAIN PATTERN.
- o LOW SIDELOBES OUTSIDE THE TACTICAL AREA.
- o FREQUENCY-HOPPING SPREAD-SPECTRUM WITH ON-BOARD PROCESSING
- o MILLIMETER-WAVE MULTI-BEAM ARRAY
- o POWER INVERSION ALGORITHM FOR HIGH RESOLUTION JAMMER NULLING.

SLIDE 2

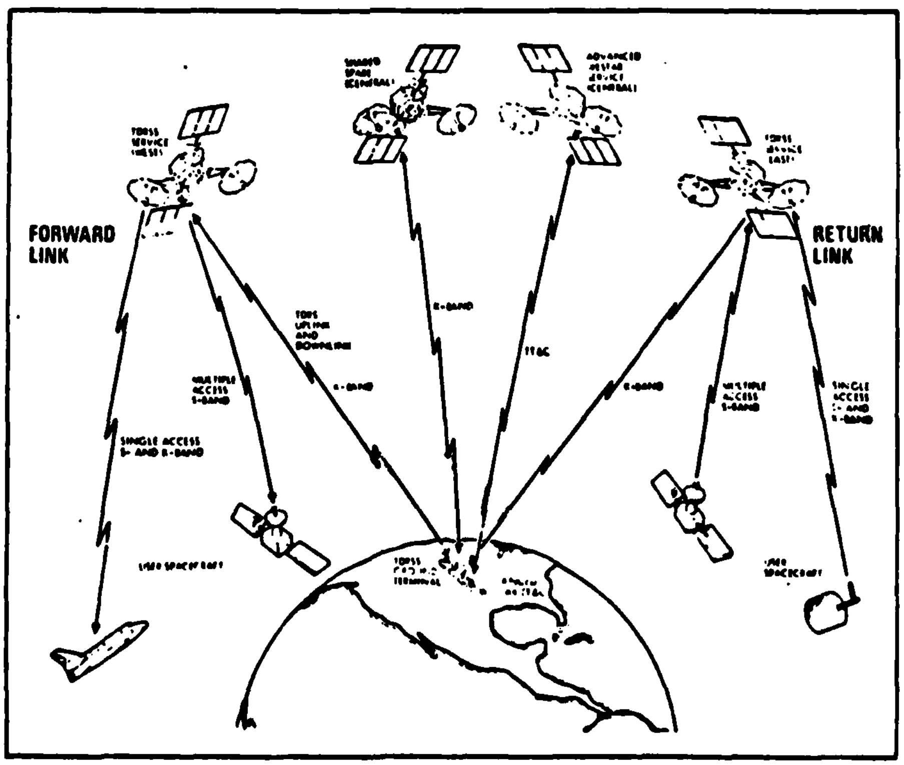
TDRSS

TRACKING AND DATA RELAY SATELLITE SYSTEM

A SYSTEM OF TRACKING AND DATA RELAY SATELLITES
IN GEO-STATIONARY ORBIT, TOGETHER WITH THE GROUND
TERMINAL AT WHITE SANDS, N.M., TO REPLACE NASA'S
GLOBAL NETWORK OF GROUND STATIONS

SLIDE 3

Figure 1. TDRSS SERVICE



SLIDE 4

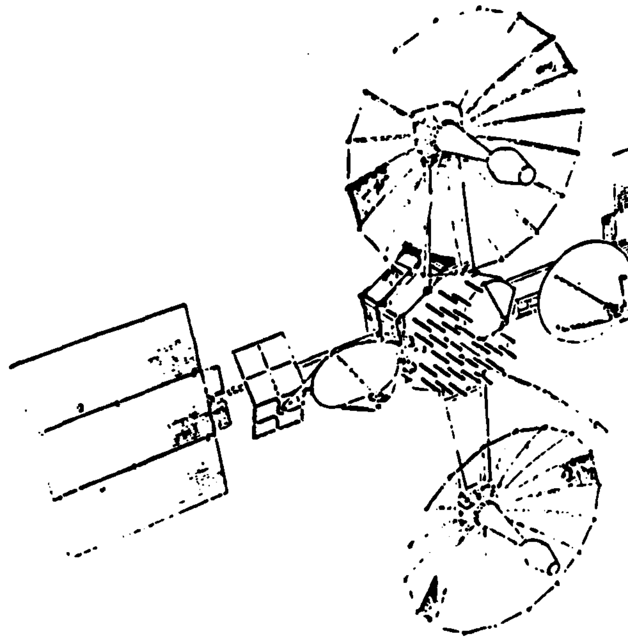


Figure 2. Spacecraft Configuration

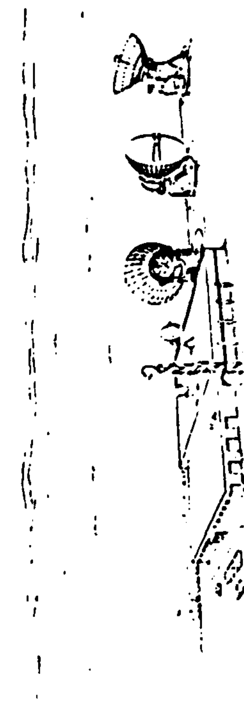
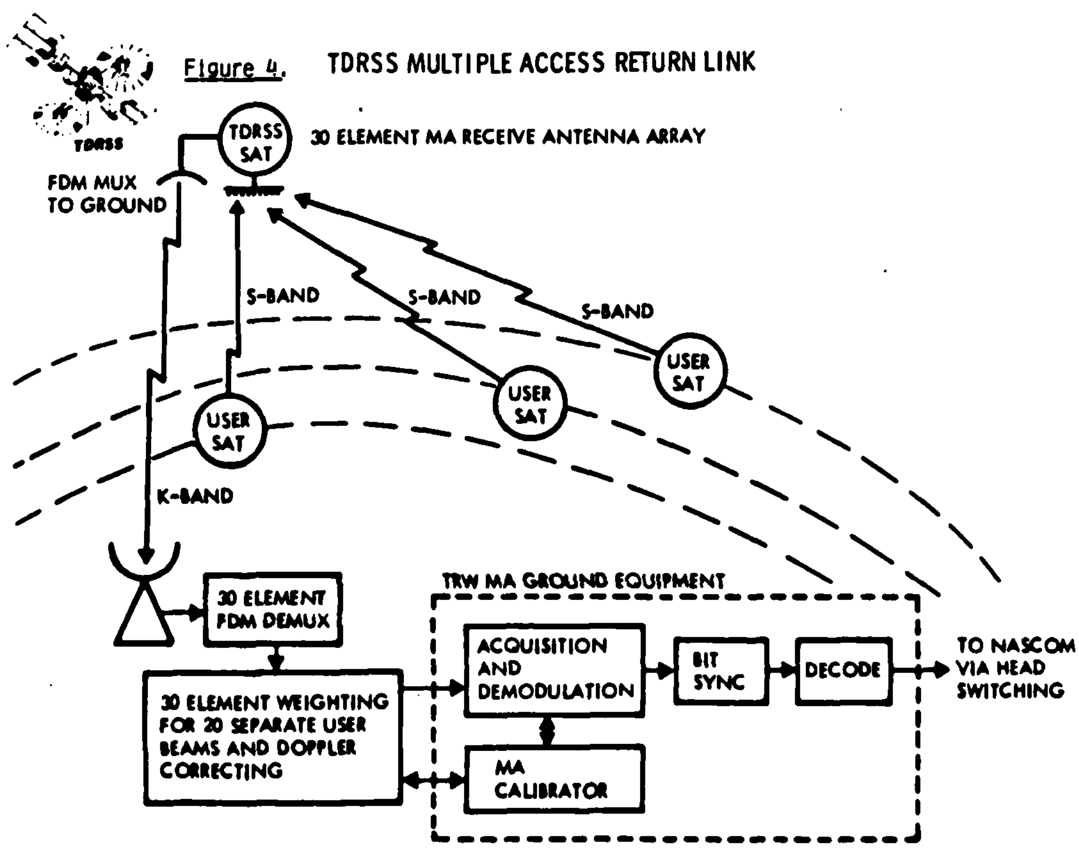


Figure 3. Ground Station at White Sands, NM

SLIDE 5

Figure 4. TDRSS MULTIPLE ACCESS RETURN LINK



SLIDE 6

(TDRSS)-Tracking & Data Relay Satellite System
(MA)-Multiple Access

-20 user satellite return link
-30 element phased array
-code division multiplexing
-remote beamforming
-adaptive calibration

SLIDE 7.

PN DIRECT SEQUENCE MODULATION

CHIP RATE: 3 MBPS
DATA RATE: 100 BPS TO 50 KBPS
FREQUENCY: 2287.5 MHZ
PERCENT BANDWIDTH: 0.26 PERCENT

- REDUCES USER FLUX DENSITY TO CCIR GUIDELINES
- PROVIDES USER IDENTIFICATION
- REDUCES MUTUAL INTERFERENCE 17.8 TO 44.8 DB
- REJECTS NARROWBAND INTERFERENCE
- PERMITS ACCURATE RANGING

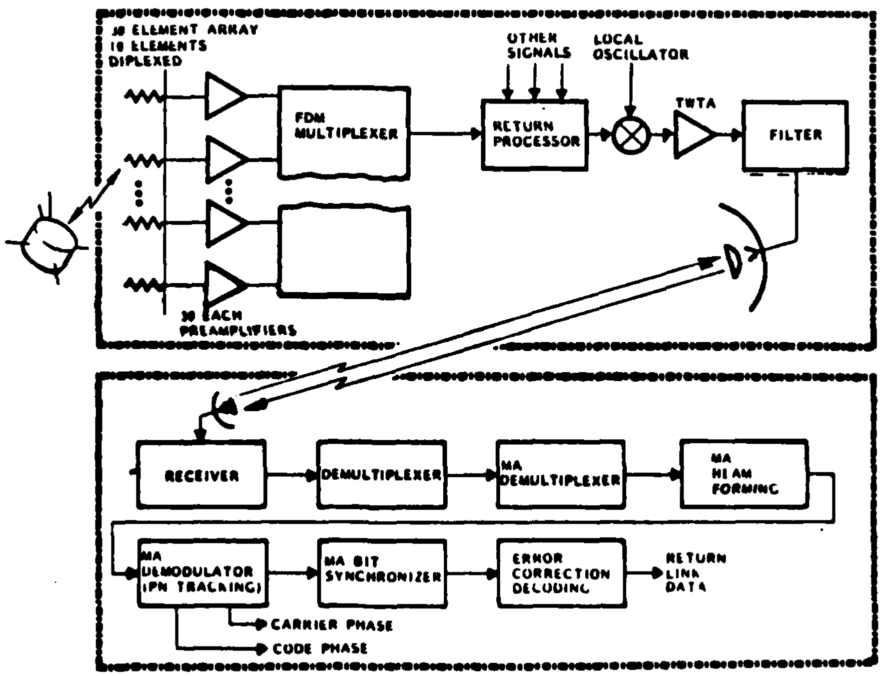
SLIDE 8

REMOTE BEAMFORMING

- 30 ELEMENT CHANNELS COHERENTLY TELEMETERED TO GROUND STATION
- 20 BEAMFORMERS EACH COMBINE 30 ELEMENT CHANNELS TO TRACK DEDICATED USERS
- REMOVES CONTROL COMPLEXITY FROM SPACECRAFT
- PILOT-TONE TECHNIQUE FOR DOWNLINK PHASE REFERENCE
- IONOSPHERIC TIME-DELAY VARIATIONS
- RANDOM PHASE AND GAIN DRIFTS OCCUR
- CALIBRATION REQUIRED

SLIDE 9

Figure 5. MA LINK PROCESSING SIMPLIFIED BLOCK DIAGRAM



SLIDE 10

THE PROBLEM

CALIBRATE AND CONTROL 20 BEAMFORMERS TO TRACK MOVING USER SATELLITES IN LOW EARTH ORBIT WITHIN 0.8 DB OF THEORETICAL SIGNAL/NOISE RATIO

SLIDE 11

CONTROL OPTIONS

- OPEN-LOOP CONTROL ONLY
- FULL-TIME ADAPTIVE CONTROL ONLY
- TIME-SHARED ADAPTIVE CONTROL ONLY
- OPEN-LOOP PREDICTION WITH TIME-SHARED ADAPTIVE CALIBRATION

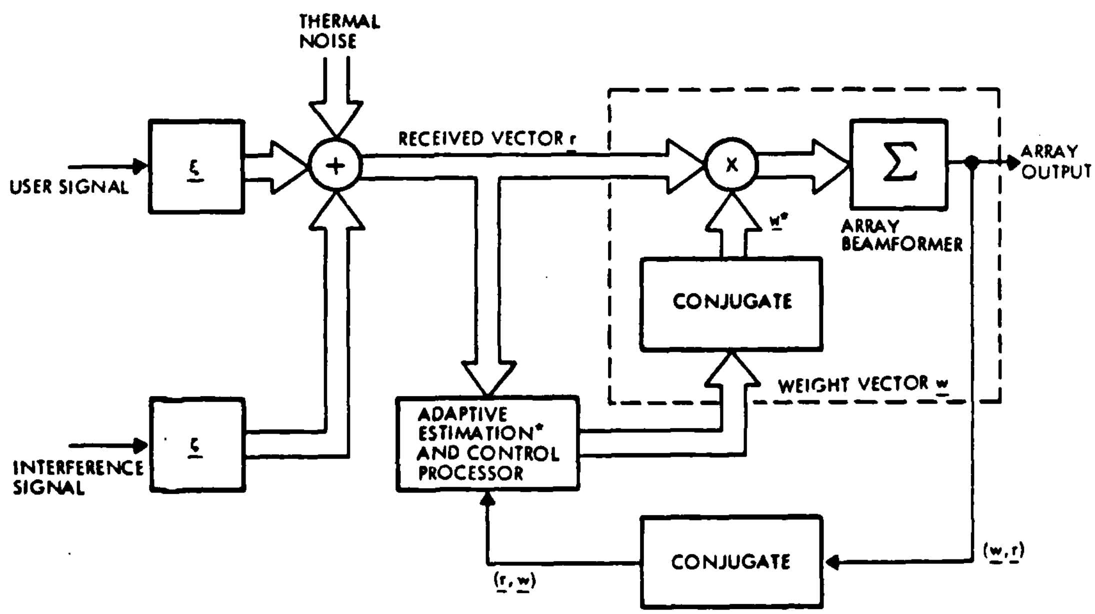
SLIDE 12

THE SOLUTION
(CONTROL AND CALIBRATION CONCEPT)

- "OPEN-LOOP" POINTING PROGRAM ESTIMATES OPTIMAL WEIGHT VECTORS FROM
  - SATELLITE EPHEMERIDES
  - ORBITAL MODEL
  - CALIBRATION VECTORS
- A CLOSED-LOOP ADAPTIVE CONTROL SYSTEM PERIODICALLY UPDATES WEIGHT VECTORS FOR A KNOWN SOURCE LOCATION
- CALIBRATION RESETS "INITIAL CONDITIONS" FOR OPEN-LOOP PREDICTION DYNAMIC MODEL

SLIDE 13

Figure 6.



*Incorporating spread-spectrum processor for signal and interference estimation

SLIDE 14

MAXIMUM ARRAY S/N

$$\frac{S}{N} = \frac{|(\underline{W}, \underline{\xi})|^2}{(\underline{W}, \Phi\underline{W})} \leq (\underline{\xi}, \Phi^{-1} \underline{\xi}) \qquad \text{(Eq. 1)}$$

MAXIMIZED BY

$$\underline{W} = \Phi^{-1} \underline{\xi} \qquad \text{(Eq. 2)}$$

Φ = SPATIAL COHERENCE MATRIX OF ARRAY SIGNALS-PLUS-NOISE

SLIDE 15

FRÉCHET DIFFERENTIAL

LET T BE A TRANSFORMATION IN AN OPEN DOMAIN D OF A NORMED SPACE X HAVING RANGE IN A NORMED SPACE Y. IF, FOR FIXED $x \in D$ AND EACH $h \in X$ THERE EXISTS A $\delta T(x;h)$ WHICH IS LINEAR AND CONTINUOUS WITH RESPECT TO h SUCH THAT

$$\lim_{\|h\| \to 0} \frac{\|T(x+h) - T(x) - \delta T(x;h)\|_Y}{\|h\|_X} = 0 \qquad \text{(Eq. 1)}$$

THEN T IS SAID TO BE FRÉCHET DIFFERENTIABLE AT x AND $\delta T(x;h)$ IS SAID TO BE THE FRÉCHET DIFFERENTIAL AT x WITH INCREMENT h.

SLIDE 16

COMPLEX GRADIENT OF SNR

THE FRECHET DIFFERENTIAL AT ω WITH INCREMENT h FOR THE RATIO OF TWO HERMITIAN FORMS

$$T(\omega) = \frac{(\omega, A\omega)}{(\omega, B\omega)}$$

FOR $\omega, h \in V^{+}$, A REAL VECTOR SPACE OF ORDERED PAIRS OF REAL VECTORS, IS

$$\delta T(\omega; h) = 2Re \frac{(A\omega - TB\omega, h)}{(\omega, B\omega)}$$

WHERE $\omega \in V^{+}$ CONSIDERED AS A COMPLEX VECTOR SPACE.

THE COMPLEX GRADIENT, WHICH MAXIMIZES THE NORM OF THE FRECHET DIFFERENTIAL, IS

$$g = A\omega - TB\omega$$

THUS, STATIONARY POINTS OF THE SNR OCCUR AT EIGENVECTORS OF THE EIGENVALUE PROBLEM

$$A\omega = \lambda B\omega$$

SLIDE 17

MAXIMUM SNR ALGORITHM

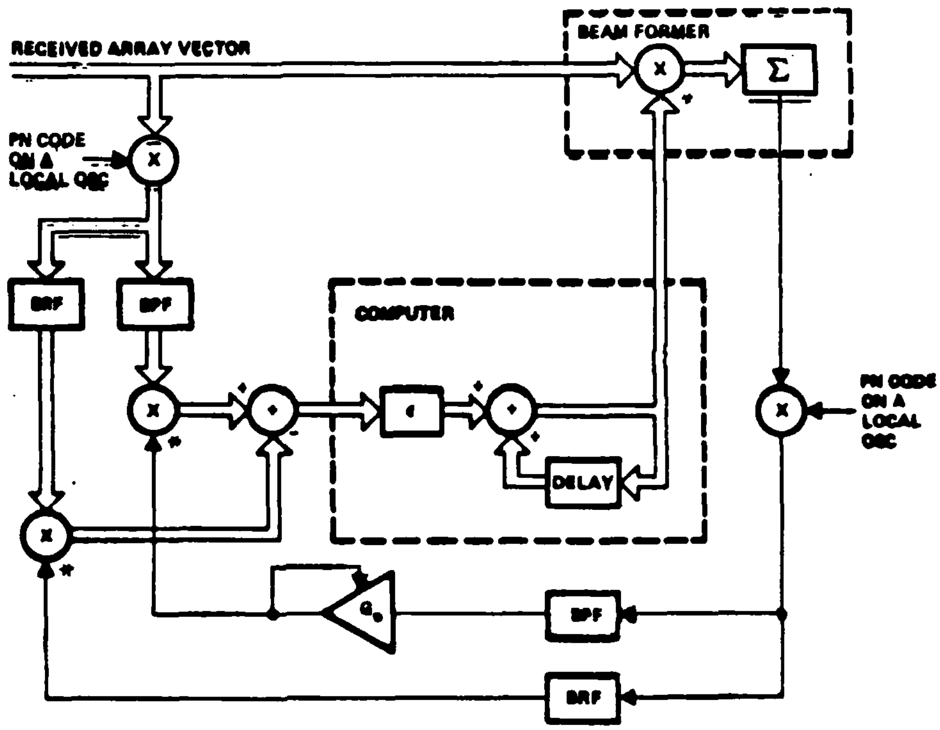
$$\omega_{n+1} = \omega_n + \mu_1 (\ell_n A\omega_n - B\omega_n) \qquad \text{(GRADIENT WEIGHT UPDATE)}$$

$$\ell_n = \frac{C_n}{(\omega_n, A\omega_n)} \qquad \text{(AGC EIGENVALUE ESTIMATE)}$$

$$C_{n+1} = C_n + \mu_2 (R - \|\omega_n\|^2) \qquad \text{(NORMALIZATION)}$$

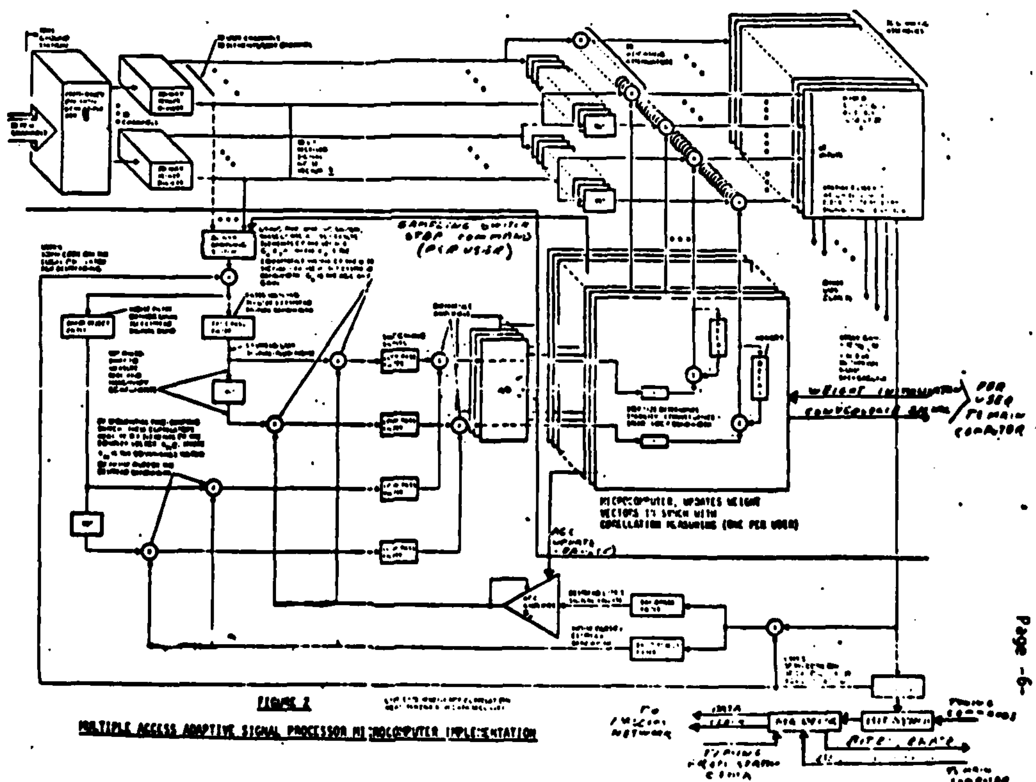
SLIDE 18.

Figure 7. **MAXIMUM S/N ARRAY PROCESSOR**



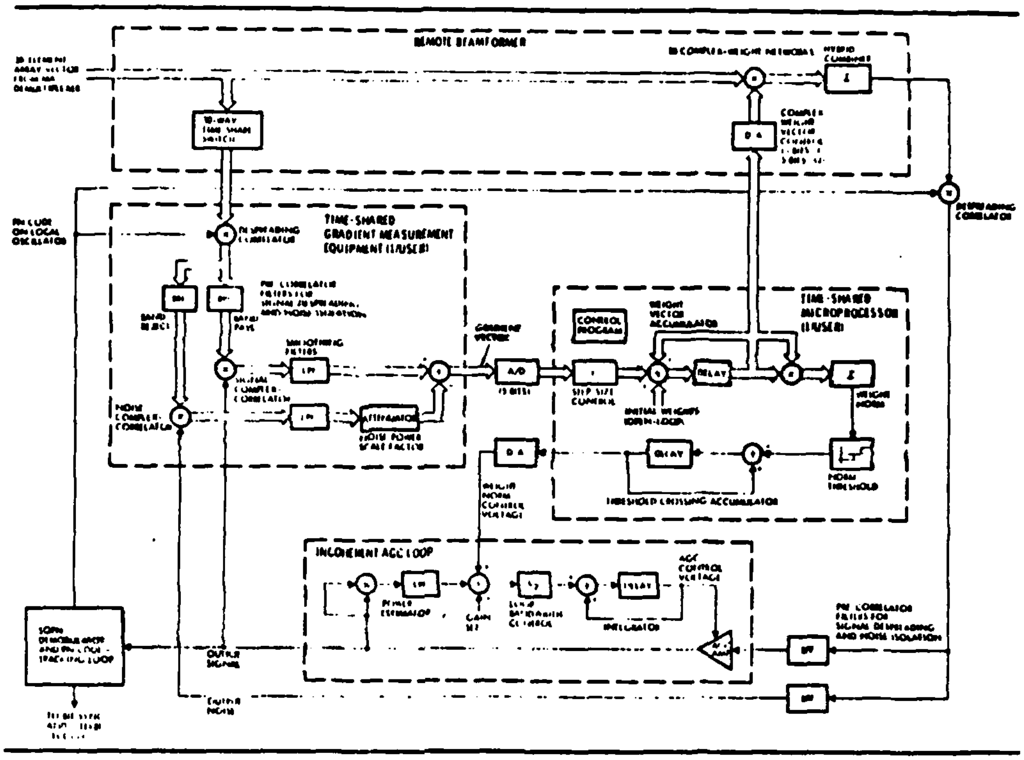
SLIDE 19

Figure 8.



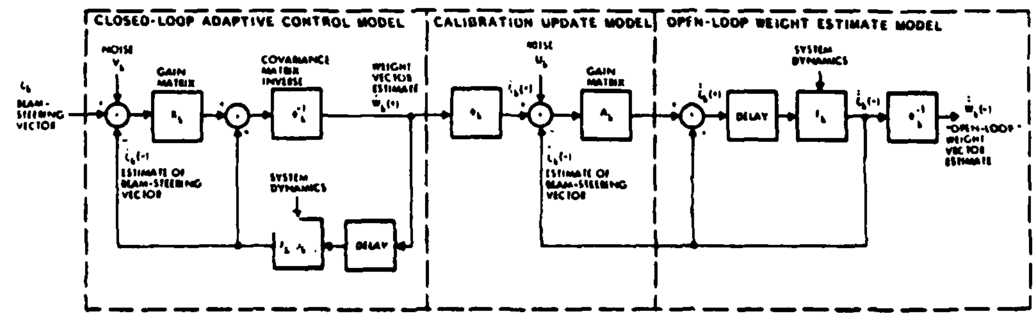
SLIDE 19a

Figure 9.



SLIDE 19b

Figure 10. CONTROL AND CALIBRATION CONCEPT



SLIDE 20

CONCLUSIONS

- USE SYSTEM DYNAMIC MODEL TO REDUCE CONTROL LOOP BANDWIDTH
- REDUCE COMPLEXITY BY TIME-SHARING ADAPTIVE CONTROL LOOP, INFERRING CALIBRATION BY COMMONALITY
- APPLICATIONS TO AVIONICS, LASER BEAM-STEERING AS WELL AS SATELLITES

SLIDE 21.

THEORETICAL RESEARCH NEEDS

o PERFORMANCE CRITERIA FOR MULTIPLE ACCESS ADAPTIVE ANTENNAS

  o CRITERIA FOR BEAM STEERING VECTOR PHASE
  o OPTIMUM ALLOCATION FOR MULTIPLE USERS
  o MULTIPLE ACCESS INFORMATION THEORY APPLICATION

o PERFORMANCE PREDICTION ANALYSIS

  o WEIGHT TOLERANCE ANALYSIS FOR NON-COMMUTING MATRICES
  o CONVERGENCE SPEED ANALYSIS

o IMPLEMENTATION

  o PROCESSING THROUGHPUT ON THE CRITICAL PATH
  o ARRAY OR NETWORK COMPUTER ARCHITECTURE BASED ON MATRIX FACTORIZATION

o OPTIMUM STRATEGIES

  o PULSED JAMMERS
  o PARTIAL BAND OR FREQUENCY FOLLOWING JAMMERS

SLIDE 22.

TECHNOLOGY NEEDS

- o LOW NOISE MILLIMETER WAVE AMPLIFIERS OVER 5% BANDWIDTH
- o 16 BIT COMPUTERS WITH 1-2 MOPS THROUGHPUT (FIXED POINT), LOW POWER (10-20 WATTS) AND RADIATION RESISTANCE = 10^6 RADS
- o MATCHING 16K FEC MEMORIES
- o DIGITAL BEAMFORMERS (NEAR FUTURE)
- o OPTICAL ADAPTIVE ALGORITHM IMPLEMENTATION (FAR FUTURE)

SLIDE 23.

APPROACH FOR PROOF OF CONVERGENCE OF MAXIMUM SNR ALGORITHM

FOR THE SINGLE-POINT SOURCE CASE,

$$\dot{w} = \mu_1((\ell-1)\xi\xi^{\dagger} - B)w = \mu_1 H(\ell)w$$

DECOMPOSE w INTO EIGENVECTORS OF H AND SHOW THAT THE SQUARED-COHERENCE OF THE DOMINANT MODE GROWS AS

$$\frac{d}{dt}\ln|\rho_1|^2 = 2\mu_1(\lambda_1 - \overline{\lambda})$$

WHERE λ_1 IS THE DOMINANT EIGENVALUE OF $H(\ell)$, $\overline{\lambda}$ IS THE MEAN EIGENVALUE $(w,Hw)/(w,w)$ AND THE LOGARITHMIC RATE OF GROWTH OF THE WEIGHT NORM-SQUARED IS $2\mu_1\overline{\lambda}$.

THE DOMINANT EIGENVECTOR IS

$$e_1 = (\lambda_1 I + B)^{-1}\underline{\xi}$$

AND APPROACHES THE OPTIMUM SOLUTION AS $|\rho_1|^2 \to 1$.

SLIDE 24.

MULTIPLE ACCESS ADAPTIVE ANTENNAS FOR COMMUNICATION SATELLITES

James E. DuPree*

ABSTRACT

Spread-spectrum coding, combined with signal and noise adaptive antennas, provides an effective way of dealing with an uncertain and even hostile communications environment. These systems are beginning to be applied in communication satellites. Two illustrations motivate the discussion of some system design choices and mathematical fundamentals which should be understood to make effective application of adaptive antennas in the new generation of multiple access satellites. Research needs, both theoretical and technological, are discussed with the object of improving the breed.

INTRODUCTION

A new generation of communication satellites is being born now. One is already in orbit, with identical versions soon to follow. And other embryos of this new generation are developing in the corporate and government laboratories. Still others are yet a "gleam in the eye" of their designers. It is to these that we address ourselves here.

This new generation will be far smarter than previous satellites. They will, of course, carry on simultaneous streams of conversation with many different users in the same frequency band independently. More significantly, they will adapt to changes in their signal environment or even failures in their components by maximizing their performance in a number of scenarios far beyond the capability of their designers to anticipate or enumerate.

Here is a report on the attributes of this new generation. Its purpose is to show, by examples and analysis, how they work, their capabilities, and the need for future development. Two examples are used to motivate the discussion with real world applications. Then, a more abstract and fundamental discussion of the adaptive process, performance criteria selection, and designing the adaptive solution is presented. This provides the basis for suggesting theoretical research topics and technological needs for future systems.

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APPLICATIONS

This discussion is motivated by two examples of the way in which adaptive antennas can be applied to communication satellites. The first example describes a part of the TDRSS satellite which was launched by the Shuttle orbiter "Challenger" last month. This is the first of a network of relay satellites which will be able to communicate with 20 low earth orbit user satellites simultaneously through a single array antenna. The second example is a hypothetical application of a nulling antenna on a military communications satellite, to provide area coverage for tactical forces operating near powerful uplink jammers.

TDRSS Multiple Access System. The Tracking and Data Relay Satellite System (TDRSS) consists of three identical satellites in synchronous orbit and a ground terminal at White Sands, NM. Two active satellites provide 85% coverage of low (200 KM) orbiting satellites, increasing to 100% coverage for 1200 KM satellite altitudes. The third satellite is an on-orbit spare. Each satellite has one S-band multiple access (MA) communication system capable of supporting up to 20 simultaneous MA return links and one MA forward link, two S-band single access (SSA) systems and two K-band single access (KSA) duplex communication systems for TDRSS communications. The Ground-to-Space and the Space-to-Ground links are at K-band through 60 foot (18.3 meter) ground station antennas, of which there are three. Figure 1 illustrates the TDRSS network concept. Figure 2 shows the relay satellite and ground station in more detail.

CCIR flux density limits constrain the TDRSS relay and user satellite maximum EIRP/Hz. Thus, frequency division multiplexing is not a practical way to provide multiple access service in this system. PN spectrum spreading with 20 unique user codes over a 5 MHz bandwidth and EIRP constraints are used to satisfy the CCIR limits and to provide a flexible multiple access service for 20 simultaneous user satellites. The data rates in this service vary from 100 bps to 50 Kbps, QPSK. The relay satellite antenna requires a net 30 dB gain to close the link budgets. The orbits of the user satellites cover a 26° cone field of view at each relay satellite. User satellites are "handed-over" from one relay satellite to the other in their orbits.

The solution to providing 30 dB of gain for each of 20 users over a 26° field of view is *remote beamforming*, using a 30 element array of 16 dB helical antennas and low-noise amplifiers to collect multiple-access signals over the FOV, combined with 20 independently controlled beamformers, which sum the element signals in the ground station. A 30 channel coherent K-band FDM downlink sends the frequency-translated element signals to the ground station for assembly with the proper phase and amplitude weighting by 20 computer-controlled beamformers. (See Figure 3).

Although user orbits can be predicted accurately from tracking data, one must initialize and periodically update open-loop computer programs which control such beamformers, to counteract phase drifts and gain variations occuring in the satellite and in the ionospheric path of the FDM downlink, so that the signals will be correctly assembled for maximum

output signal-to-noise ratio at the ground station. The initial values of the beamformer weights are obtained from a time-shared adaptive controller, which is periodically connected to each of the beamformers and adjusts the weight vector to maximize the output signal-to-noise ratio. Between "calibrations", the ground station computer predicts the dynamic behavior of the system and adjusts the beamformer weights to maintain the return link beams on each of the users. This concept is similar to that used in Kalman filters, in which the system dynamics are incorporated into the input signal prediction. (See reference 1&2 for more details).

The adaptive controller, known as the MA Calibrator, is based on a maximum SNR algorithm to be described later in this paper. Nearly all of the signal processing related to correlation and filtering in the algorithm is done digitally. The weight algorithm is implemented with an MC6800-based 8-bit microcomputer system. The firmware was programmed in assembler language and is stored in a 4K program store of EPROMs. Tables are stored in an 8K core RAM. The beamformers are digitally controlled analog devices which control the gain of each of the inphase and quadrature channels in each of each of 30 inputs to each of 20 beamformers. Thus, there are 1200 separate gains which must be orchestrated to make this system of 20 beamformers do its job of combining the array signals from any of the three satellites.

Although nulling is not necessary to provide isolation between the MA users (EIRPs are allocated according to data rate and there is adequate spread-spectrum processing gain to avoid mutual interference), the adaptive performance criterion is general enough to accomodate strong interferers. The optimum weight vector obtained by the adaptive controller produces nulls in the antenna pattern at the angles of stron interferers which might occur in the field of view. However, the calibration beacon source has a fixed location. It is desireable to remove the nulls from the pattern for open-loop steering between calibration intervals. This can be done by measuring one of the correlation vectors from which the gradient is derived, provided that the controller is in the converged state, as will be explained later. Open-loop nulling can be provided on known interference sources, using the inverse of an a-priori predicted noise covariance matrix based on known array and line-of-sight geometry.

The resulting system is an economical solution to adaptive control in which the primary mission is to counteract gain and phase shift drifts, variations in amplifier noise powers, and failures of array element channels, yet continue to operate near the theoretical maximum signal-to-noise ratio for the conditions that exist. This system will continue to operate with up to third of its elements failed. A time-shared adaptive controller is more economical than full-time adaptive control of the user beamformers. However, if strong external interference from randomly located sources were to become a problem in the future, or were needed in a future application, the same algorithm and control system could be applied to full-time adaptive control of each of the beamformers.

Tactical Area Coverage with Nulling. The second application which motivates this discussion is tactical spread-spectrum communications between units of a ground mobile force and headquarters, by means of an area coverage pattern a few hundred miles square. Hostile forces on the periphery of this area may attempt to disrupt communication by jamming the satellite uplink, using broad-band pulsed signals over the spread-spectrum bandwidth. The pattern of user traffic in this application demands a relatively flat gain over the specified area, but with high resolution nulling capability anywhere within the coverage area or on the near-in sidelobes. The sidelobes far from the coverage area must remain low enough to reject any reasonable jammer threat without adaptive nulling. If the coverage area is a relatively small portion of the earth, which can be covered by contiguous beams, the preferred type of antenna is most likely an offset-fed Cassegrain antenna with multiple feed horns and a combining network and control system on-board the satellite. The nulling algorithm would be designed to null strong interference, but otherwise to maintain the pattern in a least-squares approximation to the designated coverage pattern, which would meet minimum G/T specifications within the coverage area. The reflector size and number of elements would be selected to meet the resolution and coverage requirements. No specific design parameters are advanced here, but economic factors favor application of this type of antenna at the shorter millimeter wavelengths. Global deployment of such a tactical area coverage beam would be obtained by gimballing the small beam antenna to specified service areas.

By virtue of a large amount of spread-spectrum processing gain which would be available at the higher frequencies, there would be a large difference in the user and jammer powers. The jammer is forced to spread his power over the entire bandwidth, assuming that a signal processing strategy is designed to defeat frequency following jammers. Thus, to be effective, the jammer must have a wide bandwidth and a large EIRP. Then, the power inversion behavior of certain adaptive algorithms will discriminate against the jammer because of his large power level. User signals will not be nulled because they are significantly below the noise level in the full spread-spectrum bandwidth.

The quiescent beam pattern can be specified in this system by a beam steering vector which sets the weights, in the absence of nulling, to form a least-squares approximation to a specified field pattern, over the coverage area. The solution takes into account the cross-coupling of beam-response functions. One of the outstanding theoretical problems with this type of antenna systems concerns the optimum phase of the field pattern. Since phase information is eliminated in the power gain specification of the pattern and since relative phase is of no concern to a set of multiple access users whose signals are not phase coherent, it would seem that the phase can be arbitrary. However, as shown by Potts, Mayhan, and Simmons (Ref. 3), experimentally determined phase gradients in the beam-steering vector affect the angular resolution in an adaptive nulling antenna. It remains to state the approach in a performance criterion.

THE ADAPTIVE PROCESS

Adaptive problem solving involves a dual relation with the world--a state description and a process description, to paraphrase Herb Simon (Ref. 4). The first describes a situation. The second provides a procedure for generating the situation. Algorithms are the process description, as a set of rules for optimizing the performance criterion with a digital or analog machine.

Performance Criterion. Selection of performance criteria involves the designer in means-ends analysis. Not only must the criterion improve the performance of the antenna as an information channel, but a mathematical base of support must be present to permit the performance features to be predicted and to provide the process description. A technological base of support must be present to realize the machine.

The performance criteria in current use are real scalar functionals of the weight vector. Whether the vector space is treated as real or complex depends upon the mathematical basis for the operations to be performed, as we shall see later. Constraints are included by the method of Lagrange multipliers. But one must find a single descriptor of performance which can be expressed as a functional of the weight vector for the array. In the following paragraphs, we examine some candidate criteria which might be applied to the example systems.

The *least-squared error* is historically and practically important in systems where a desired output signal is known or can be generated by some process to serve as a reference. The element signals of the array antenna could be combined to approximate the desired signal and the optimum weight vector would minimize the mean squared error. There is a vast literature on least-squares estimation theory. This theory has been applied by Widrow, *et. al.* to a wide variety of adaptive antennas and adaptive filters in a stochastic approximation approach called the LMS algorithm (Refs. 5-7).

To illustrate the least-squares criterion, represent the desired signal by d(t) and the array signal-plus-noise vector by x(t). The beamformer weight vector conjugate, $\underline{w}^*$, produces an array output mean-square error of

$$\overline{|\varepsilon|^2} = \overline{|d(t) - \underline{w}^{\dagger}x(t)|^2}$$

$$= \overline{|d|^2} - 2Re(\underline{w},\underline{r}_{xd^*}) + (\underline{w},R_{\underline{xx}}\underline{w}) \qquad (1)$$

where

$$\underline{r}_{xd^*} = \overline{\underline{x}d^*} \qquad \text{and } R_{\underline{xx}} = \overline{\underline{x}\underline{x}^{\dagger}} \qquad (2)$$

denote the cross-correlation vector and the covariance matrix, respectively. The overbar denotes the mathematical expectation and it is assumed that the processes are stationary. The optimum weight vector is given by

$$\underline{w}_{opt} = \underline{R}_{xx}^{-1} \, \underline{r}_{xd^*} \tag{3}$$

The least-squared error criterion is not well-suited to either of the example systems, however. In the TDRSS application, beam-steering vector estimates must be derived from the adaptive control system to initialize the open-loop pointing program. The desired signal is not present to use as a reference. Although the reference signal could be synthesized by despread/filter/respread techniques, as done by Riegler and Compton (Ref. 8), the signal reference could not be easily integrated into an open-loop steered array. In the area-coverage beam example, the signal has a known direction of arrival, namely from within the specified area, but the desired signal is entirely unknown. The user signals tend to occur in short bursts, so that adaptation to individual user signals is not practical.

The <u>maximum signal-to-noise ratio</u> is an important criterion because the information transfer rate is directly related to the signal-to-noise ratio. This criterion is sometimes associated with the Howells-Applebaum algorithm (Refs. 9,10). This algorithm solves one case of the maximum signal-to-noise ratio problem, for a signal with a known direction of arrival. The direction of arrival is described by a beam steering vector.

Let $\underline{\xi}$ denote the beam-steering vector for a known direction of arrival of a user signal of power, S, in a noise environment described by the covariance matrix, R_{nn}. The array output SNR produced by a conjugate vector, $\underline{w}^*$, is

$$\mathrm{SNR} = \frac{S(\underline{w}, \underline{\xi}\underline{\xi}^{\dagger}\underline{w})}{(\underline{w}, R_{nn}\underline{w})} \tag{4}$$

The maximizing weight vector, $\underline{w}_{opt}$, is given by Schwarz's inequality as

$$\underline{w}_{opt} = (\text{constant}) R_{nn}^{-1} \underline{\xi} \tag{5}$$

It was shown by Griffiths (Ref. 11) that, for this special case, both the least-squared error criterion and the maximum signal to noise ratio criterion produce the same output signal to noise ratio. But they do not produce the same mean-squared error.

In the radar sidelobe cancellation system which motivated the development of the Howells-Applebaum algorithm, the beam steering vector is often known. The desired beam is a single-mode, high resolution, pencil-shaped pattern in the transmit direction. The beam steering vector can be calculated from the array geometry and the known direction of arrival.

The direction of arrival can not be used as a criterion in TDRSS because of the uncertainty due to phase drifts in the satellite and ionos-

spheric path. The beam steering vector estimate is a major objective of the adaptive control system design. It is shown below that a more general maximum SNR criterion yields a useful algorithm for the TDRSS application.

Known direction of arrival can be assumed for the area coverage beam system. But, as described previously, there is no known criterion for optimally selecting the phase relation between the elemental beams which comprise the area coverage beam.

A *general maximum SNR* criterion assumes that the direction of arrival is unknown. The beamformer output SNR is the ratio of two Hermitian forms

$$SNR = \frac{(\underline{w}, R_{ss}\underline{w})}{(\underline{w}, R_{nn}\underline{w})} \tag{6}$$

where $\underline{w}$ is a complex weight vector, R_{ss} is the signal covariance matrix, and R_{nn} is the noise covariance matrix, including thermal noise, which makes it positive definite. The optimum weight vector satisfies the eigenvalue problem

$$R_{ss}\underline{w} = \lambda R_{nn}\underline{w} \tag{7}$$

where λ is the output SNR. For weights which satisfy this problem, the SNR has a discrete eigenvalue spectrum. If the eigenvalues are distinct, the maximum SNR eigenvector is unique, to within a scalar constant factor. The vectors on either side of the eigenvalue problem are the desired beam steering vectors. In the single point-source user signal case, the maximum eigenvector solution is the same as that obtained by the Howells-Applebaum algorithm approach and the least-squared error method. The differencs lies primarily in the implementation of the algorithm.

The mathematical base of support for maximum SNR criteria is in the theory of Hermitian forms. Because of their importance in quantum mechanics, there is a large body of mathematical physics literature concerning Hermitian forms. Eigenvalue problems are a fundamental part of this literature. Computational algorithms exist for solving these eigenvalue prolems.

It is known that if the largest eigenvalue of (7) is distinct, then a computational algorithm exists which will determine the eigenvector uniquely, to within a scalar constant. However, if the eigenvalue problem has two or more solutions for the same value of λ, the corresponding eigenvectors are not uniquely determined by any computer algorithm. Thus, maximum SNR is not a useful criterion when the largest eigenvalue is degenerate. The solution will not be uniquely found.

In the area coverage example, the maximum eigenvalue is degenerate. Clearly, maximizing the SNR is not a useful criterion, for this case. In order to obtain a more rigorous performance criterion, one may use the

mathematical structure and formalisms of quantum mechanics and information theory to analyze the antenna as an information channel. In adopting the formalism of quantum mechanics, we make no claims to physical interpretation of the antenna as a quantum device.

Let the state of the antenna be described by the weight vector, $\underline{w}$. Obserables are described by matrix operators; in this case, either the covariance matrix or the cross-power spectral density matrix. The Hermitian forms represent expected values of the observable, the output power. The signal-to-noise ratio is the ratio of the expected value of the signal power and the expected value of the thermal noise and jammer power.

Since user signals are spread over the coverage area and are not coherent, take the signal covariance model to be an identity matrix,

$$R_{ss} = I \tag{8}$$

The identity matrix can be resolved by a set of projection operators

$$I = \sum^{N} \underline{\xi}_i \underline{\xi}_i^{\dagger} \tag{9}$$

where $\underline{\xi}_i$ is a normalized eigenvector of the noise covariance matrix.

$$R_{nn}\underline{\xi}_i = \mu_i \underline{\xi}_i \tag{10}$$

Degenerate eigenvalues of R_{nn} would occur for the case of equal thermal noise powers in different elements.

Express the weight vector as a Fourier series expansion in the eigenvectors of the noise covariance matrix.

$$\underline{w} = \sum^{N} c_i \underline{\xi}_i \tag{11}$$

and the corresponding "average SNR" at the output of the array is

$$\langle\psi\rangle = \frac{(\underline{w},\underline{w})}{(\underline{w},R_{nn}\underline{w})} = \sum^{N} p_i \psi_i \tag{12}$$

where p_i is a probability for the antenna in the $i\underline{th}$ mode and ψ_i is the SNR in this mode. These quantities are related to c_i by the relations

$$p_i\psi_i = \frac{|c_i|^2}{\sum^N |c_i|^2\mu_i} \qquad (13)$$

and

$$\psi_i = 1/\mu_i \qquad (14)$$

Considering the antenna as an information channel, what is the effect of a given $\underline{w}$ on the information rate? In a given eigenvector state, the information transfer rate is (Ref. 12)

$$R_i = \frac{1}{4\pi}\int_\Omega \log(1 + \psi_i)d\omega \qquad (15)$$

where the SNR in the i<u>th</u> state, ψ_i, is computed as a function of frequency using the cross-power spectral density matrices for the signals and noise. Then, the average rate of information transfer is bounded by

$$\overline{R} \leq \frac{1}{4\pi}\int_\Omega \log(1 + \langle\psi\rangle)d\omega \qquad (16)$$

When there are degenerate eigenvalues, there are many ways to achieve the same average SNR, hence the same average information transfer rate. We seek a fair allocation of information rates among the users. such that a fixed average information rate,

$$\overline{R} = \sum^N p_i R_i \qquad (17)$$

is maintained.

In the absence of other information constraining the allocation, it seems reasonable to use the method of maximum entropy; i.e., choose the distribution of the set of probabilities, $\{p_i\}$, to maximize the entropy,

$$H = -\sum^N p_i \log(p_i) \qquad (18)$$

subject to a fixed average information rate (17), and normalized probability

$$\sum^N p_i = 1 \qquad (19)$$

By classical methods (Ref. 13), it can be shown that the probability distribution is

$$p_i = \frac{\exp(\mu R_i)}{\sum^{N} \exp(\mu R_i)} \qquad (20)$$

where the parameter, μ, is chosen to satisfy the constraint (17). A similar result is obtained if the constraint is taken to be a fixed average SNR.

From the optimum probability distribution, (20), we see that if the signal sources in each eigenvector beam have equal SNR and bandwidth, the probability distribution is uniform. However, where the SNR is low, the information transfer rate in the ith eigenstate would also be low and the probability is weighted correspondingly lower by the exponential function. Degenerate eigenvectors corresponding to the same information rate are assigned the same probability. This an intuitively reasonable result.

However, note by (13) that the probability distribution carries no phase information. Although we can determine the relative amounts of each eigenvector component in a given weight vector, we can not assign a weight vector to the antenna, since we do not know the phase relation between these components. By this, we conclude that the signal and noise description is not rich enough to describe the antenna pattern at intermediate points in a dense grid of points which would constrain the antenna pattern to unique results. A similar situation exists in quantum mechanics for systems in mixed states.

Another criterion that produces the Howells-Applebaum algorithm upon differentiation is *Noise Minimization, subject to a quadratic pattern constraint*. That is, define the performance criterion as

$$\min_{\underline{w}} \; (\underline{w}, R_{nn}\underline{w}) + \lambda \| \underline{\xi} - \underline{w} \|^2 \qquad (21)$$

Here, the Lagrange multiplier acts as a penalty function, so that the weight vector $\underline{w}$ forms a least-squares approximation to the desired pattern's beam steering vector, $\underline{\xi}$, in the absence of jammers, and is constrained from departing too much from it in the presence of jammers. But large jammers will cause a large reduction in gain in their directions. This criterion suffers the same incomplete definition as the others in the case of the area coverage beam.

The Process Description. Gradient processes are most commonly used to update the weight vector from a given value toward a value which improves the performance criterion. Because of the conjugate symmetry inherent in the array geometry, it is convenient to analyze the antenna performance criteria as real functionals of a complex vector space. This requires a close look at the mathematical fundamentals to be sure the proper definitions are available.

The complex gradient for SNR is developed in the next few paragraphs. Complex analytic signals have obvious advantages in analyzing communcation systems. The complex signal leads naturally to relations between complex transfer functions, based on the fact that the exponential function is an eigenfunction for linear, time-invariant systems. In the antenna, complex

eigenfunction for linear, time-invariant systems. In the antenna, complex weights describe the gain of the path from an array element to the beamformer output. The weight vector describes the state of the antenna. The covariance matrices of array signals are Hermitian matrices. Thus, it is natural to cast the array optimization problem in a complex vector space.

Note that the performance criteria are usually real functionals of a complex vector. The usual definition of derivative fails to work for the function $|z|^2$, where z is a complex variable, because the limits do not exist. Similar results are obtained with more advanced definitions, such as the Gâteaux and Fréchet differentials (Ref. 14).

The complex gradient can be derived from the Fréchet differential by using the dual identity of a complex vector space as a real vector space of ordered pairs of real vectors, as outlined below.

Suppose V_R is a real vector space, and let V^+ be the set of ordered pairs $\{\underline{x},\underline{y}\}$ with both $\underline{x}$ and $\underline{y}$ in V_R. Then V^+ is a real vector space under the addition rule defined by

$$\{\underline{x}_1,\underline{y}_1\} + \{\underline{x}_2,\underline{y}_2\} = \{\underline{x}_1+\underline{x}_2,\ \underline{y}_1+\underline{y}_2\} \tag{22}$$

and the rule for scalar multiplication by a real scalar, α, defined by

$$\alpha\{\underline{x},\underline{y}\} = \{\alpha\underline{x},\alpha\underline{y}\} \tag{23}$$

But the elements of V^+ also form a complex vector space under the same addition rule and the rule for scalar multiplication by a complex scalar $\alpha + i\beta$ defined by

$$(\alpha+i\beta)\{\underline{x},\underline{y}\} = \{\alpha x-\beta y,\alpha x+\beta y\} \tag{24}$$

Thus, the elements of V^+ have a dual identity; i.e., elements both of a complex vector space and of a real vector space. The rules for scalar multiplication depend upon the role assigned to V^+. This role depends upon the scalar field. The rule (24) for a complex field is not defined as scalar multiplication in a real vector space, since complex numbers are not contained in the field of real numbers.

There is a way to obtain the result of complex scalar multiplication in a real vector space, however. One may use the isomorphism between scalar multiplication by a complex number $(\alpha+i\beta)$ and matrix transformations

$$\begin{pmatrix} \alpha I & -\beta I \\ \beta I & \alpha I \end{pmatrix} \begin{pmatrix} \underline{x} \\ \underline{y} \end{pmatrix} = \begin{pmatrix} \alpha\underline{x} - \beta\underline{y} \\ \beta\underline{x} + \alpha\underline{y} \end{pmatrix} \tag{25}$$

We can show that the inner product between two elements of V^+ as a real vector space is the real part of the inner product between these two elements of V^+ as a real vector space. This is the key to identifying the complex gradient from the Fréchet differential. The inner product between two elements of a vector space V^+ is a bilinear functional which maps the vector space $V^+ \times V^+$ into the field F, and which satisfies the axioms

$$1. \quad (\underline{x},\underline{y}) = (\underline{y},\underline{x})^* \tag{26}$$

$$2. \quad (\alpha\underline{x}_1+\gamma\underline{x}_2,\underline{y}) = \alpha^*(\underline{x}_1,\underline{y}_1) + \gamma^*(\underline{x}_2,\underline{y}_2) \tag{27}$$

$$3. \quad (\underline{x},\underline{x})\geq 0 \;\forall\; \underline{x} \in V^+;\; (\underline{x},\underline{x})=0 \text{ iff } \underline{x}=0 \tag{28}$$

For V^+ as a real vector space, the inner product is $(\underline{x}_1,\underline{x}_2) + (\underline{y}_1,\underline{y}_2)$. For V^+ as a complex vector space, the inner product is defined by $(\underline{x}_1+i\underline{y}_1,\underline{x}_2+i\underline{y}_2) = (\underline{x}_1,\underline{x}_2) + (\underline{y}_1,\underline{y}_2) - i((\underline{x}_1,\underline{y}_2)-(\underline{y}_1,\underline{x}_2))$ By comparing the real part of the complex inner product with the real inner product, the assertion is proved.

Now, we can use the Fréchet differential to define the gradient of a real functional of a complex weight vector, from the dual role of V^+. Let T be a transformation in an open domain $\mathcal{D}$ of a normed space X having a range in a normed space Y. If for fixed $\underline{x} \in \mathcal{D}$ and each $\underline{h} \in X$ there exists a $\delta T(\underline{x};\underline{h}) \in Y$ which is linear and continuous with respect to $\underline{h}$ such that

$$\lim_{\|\underline{h}\|\to 0} \frac{\|T(\underline{x}+\underline{h})-T(\underline{x}) - \delta T(\underline{x};\underline{h})\|_Y}{\|\underline{h}\|_X} = 0 \tag{29}$$

then T is said to be Fréchet differentiable at $\underline{x}$ and $\delta T(\underline{x};\underline{h})$ is said to be the Fréchet differential at $\underline{x}$ with increment $\underline{h}$.

When the above definition is applied to an Hermitian form, we see the $T(\underline{z}) = (\underline{z},H\underline{z})$ is Fréchet differentiable at $\underline{z}$ for $\underline{z} \in V^+$ as a real vector space of ordered pairs of real vectors. It is not differentiable for $\underline{z} \in V^+$ as a complex vector space, because $\delta T(\underline{z};\underline{h})$ is not linear under the rule of scalar multiplication that admits complex scalars.

The Frechet differential

$$\delta T(\underline{z};\underline{h}) = (\underline{h},H\underline{z}) + (H\underline{z},\underline{h}) \tag{30}$$

is a linear functional from $V_R \to R$. The norm of this mapping is

$$\|\delta T\| = \sup_{\substack{\underline{h} \in V_R \\ \underline{h} \neq 0}} \frac{T(\underline{z};\underline{h})}{\|\underline{h}\|} \tag{31}$$

The gradient direction is defined as a unit vector, $\hat{h}$, aligned with increment $\underline{h}$ which maximizes the norm, $\|\delta T\|$. For $T(\underline{z}) = (\underline{z}, H\underline{z})$ in our example, one can show by Schwarz's inequality that the gradient direction is

$$\hat{\underline{g}} = \frac{H\underline{z}}{\|H\underline{z}\|} \tag{32}$$

and the corresponding Fréchet differential value is $2\|H\underline{z}\|$, from (30), when $\underline{h}$ is a unit vector in the gradient direction.

Similarly, it can be shown that the Frechet differential for the ratio of two Hermitian forms,

$$T(\underline{w}) = \frac{(\underline{w}, A\underline{w})}{(\underline{w}, B\underline{w})} = \lambda \tag{33}$$

is

$$\delta T(\underline{w};\underline{h}) = 2Re\frac{(A\underline{w} - \lambda B\underline{w}, \underline{h})}{(\underline{w}, B\underline{w})} \tag{34}$$

and the complex gradient can be taken as a vector aligned with

$$\underline{g} = A\underline{w} - \lambda B\underline{w} \tag{35}$$

At the stationary points of the SNR, the gradient is zero. Thus, the stationary points occur at points in the complex vector space where $\underline{w}$ satisfies the eigenvalue problem (7).

As a final example, the gradient of the performance criterion in (21) is given by

$$\underline{g} = R_{nn}\underline{w} + \lambda\underline{w} - \lambda\underline{\xi} \tag{36}$$

The algorithm which completes the process description is usually a linear update of the weight vector along the gradient direction, with the step-size determined by a compromise between convergence speed and steady-state accuracy (Ref. 14). In stochastic approximation algorithms, the gradient measurement noise effects on convergence rate and accuracy are an important consideration. However, we will not have room to discuss the stochastic properties here. See the above reference for a thorough discussion. We will present an example of the construction of an algorithm for maximum SNR, using the gradient derived in (35), and will discuss its deterministic convergence properties.

A linear weight update equation, along the gradient of the SNR, is

$$\underline{w}_{n+1} = \underline{w}_n + \varepsilon_n(A\underline{w}_n - \lambda_n B\underline{w}_n) \tag{37}$$

where we let A and B be the signal and noise covariance matrices, respectively, for convenience in notation. Note that

$$(\underline{w}_n, A\underline{w}_n - \lambda_n B\underline{w}_n) = 0 \tag{38}$$

so the gradient is orthogonal to the gradient of the norm-squared of the weight vector. Movement along the SNR gradient is tangent to the iso-contour of the weight vector.

Though we do not explicitly analyze the stochastic effects of gradient measurement noise here, it is useful to gradually decrease the step size as the optimum is reached. This is equivalent to reducing the noise bandwidth of the algorithm as a filter. To obtain the effect of gradually decreasing the step-size automatically, choose the step size factor ϵ_n as

$$\epsilon_n = \frac{\mu_1}{\lambda_n} \tag{39}$$

Also, as a practical matter, we shall want to normalize the weight vector since the variations in the weight norm will affect the gain of the beamformer and introduce excess quantization error from the digitally controlled weights. Finally, the eigenvalue can be estimated by means of an AGC amplifier in the signal path. Thus, the algorithm for maximum SNR is given by the three equations

$$\underline{w}_{n+1} = \underline{w}_n + \mu_1(\ell_n A\underline{w}_n - B\underline{w}_n) \text{ (weight update)} \tag{40}$$

$$\ell_n = \frac{C_n}{(\underline{w}_n, A\underline{w}_n)} \quad \text{(AGC gain)} \tag{41}$$

$$C_{n+1} = C_n + \mu_2(R - (\underline{w}_n, \underline{w}_n)) \text{ (normalization)} \tag{42}$$

The convergence behavior of the maximum SNR algorithm can be demonstrated for the single point source signal case. The discussion deals with equivalent differential equations. The approach used is to transform the weight update differential equation to one describing orthogonal coordinates, the eigenvectors of the matrix

$$H(\ell) = (\ell-1)S\underline{\xi}\underline{\xi}^{\dagger} - B \tag{43}$$

which partition the vector space into orthogonal modes. The weight norm-squared is fixed, and the "energy" in the weights is partitioned among these orthogonal modes. Define a coherence coefficient whose square is a measure of the proportion of the fixed "energy" which lies in the dominant mode. Because ℓ is variable, the coordinates shift direction as the algorithm converges. But the dominant mode aligns itself with the optimum weight vector as the proportion of the "energy" in the dominant mode approaches unity. If the weight vector is normalized, the coherence of the dominant mode grows in a exponential saturation curve toward unity.

Now, to demonstrate this behavior, represent the weight update differential equation as

$$\dot{\underline{w}} = \mu_1 H(\ell)\underline{w} \tag{44}$$

and let

$$\underline{w} = M\underline{y} \tag{45}$$

where M is a diagonalizing unitary transformation such that

$$M^{\dagger} H M = \Lambda \tag{46}$$

is a diagonal matrix and the norm is preserved: $(\underline{w},\underline{w}) = (\underline{y},y)$. If the commutator

$$\underline{\xi\xi}^{\dagger} B - B\, \underline{\xi\xi}^{\dagger} = 0 \tag{47}$$

then the coordinates are eigenvectors of B. Otherwise, the dominant eigenvector coordinate is

$$\underline{m}^{1} = (\lambda_1 I + B)^{-1}\underline{\xi} \tag{48}$$

and the other coordinates are in the null space of $\underline{m}^{1}$.

Thus, the transformed weight update equation is

$$\dot{\underline{y}} = \mu_1 \Lambda \underline{y} \tag{49}$$

and the norm-squared varies according to the equation

$$\frac{d}{dt}\ell n(\underline{y},\underline{y}) = 2\mu_1 \overline{\lambda} \tag{50}$$

The weights have constant norm-squared iff $\overline{\lambda} = 0$.

Define the complex coherence of the i<u>th</u> mode by

$$\rho_i = \frac{y_i}{(y,y)} \tag{51}$$

Then the growth equation for the dominant mode coherence-squared is

$$\frac{d}{dt}\ln|\rho_1|^2 = \frac{d}{dt}\ln|y_1|^2 - \frac{d}{dt}\ln(\underline{y},\underline{y}) \tag{52}$$

$$= 2\mu_1(\lambda_1 - \overline{\lambda}) \tag{53}$$

From (53), there is a monotonic growth in coherence for $\lambda_1 \geq \overline{\lambda}$. If the norm is held constant, then this condition is satisfied, because λ_1 is the dominant eigenvector of H. When λ_1=0, the dominant mode is aligned with the optimum solution and the other nodes decay.

For example, when B= σ^2I, the coherence-squared of the dominant mode grows as

$$|\rho_1|^2 = 1 - (1 - |\rho(0)|^2\exp(-2\mu_1\sigma^2 t) \tag{54}$$

Note that constant norm-squared is merely a sufficient condition for monotonic convergence. It may be that faster convergence may be obtained by controlling the gain, ℓ, in some optimal fashion. This is an interesting topic which could possibly be answered by some techniques of nonlinear optimal control.

Implementation of the maximum SNR algorithm is illustrated in simplified fashion in Figure 4. The received array vector elements are combined into a single output with a beamformer. Mathematically, this is a complex inner product. By the definition of complex inner product (26)-(27), one of the vectors is conjugated. Here, it is the weight vector. Element channel and sum channel signals are despread, using a PN code on a local oscillator. The user signal is thereby compressed into a narrow-band bandpass filter; whereas, the noise and interference signals are not correlated with the PN code and are not despread. A pair of complex correlators combine the array signal vectors with the signals processed by the beamformer (inner product) to produce the signal and noise correlation vector estimates. The gain of the signal channel is AGC controlled. The AGC amplifier serves the purpose of estimating the eigenvalue, when its output power reference is adjusted to control the weight normalization, as in (27). The difference between the two correlation vector estimates is the gradient vector estimate. This estimate is scaled by a step-size factor and added to the previous weight vector to produce the weight update. Finally, the weights are conjugated and applied to the beamformer. This convention of conjugating the beamformer requires that the complex correlators be designed in such a way that the sum inputs to the correlators are also conjugated. The weight normalization loop is not shown in this diagram. It is implemented by accumulating the sum of the squares of the inphase and quadrature weights and comparing with an internally stored normalization reference. The scaled difference updates the output power reference for the AGC amplifier in the signal correlation feedback loop.

The 30 element sample channels must usually be multiplexed into a single channel before despreading and correlation, in order to reduce the complexity. Thus, the components of the gradient vector are not measured simultaneously, but sequentially. The convergence rate does not seem to

be sensitive to whether the updates are made sequentially or in a batch. It is most convenient to process sequential updates through the microcomputer, which accumulates the weight updates controls the multiplex switch for the element sample channel, commands the AGC power reference, and controls the beamformer weights.

The signal processing, after the despreaders, is all done digitally in the TDRSS implementation. The output of the despreaders is quantized to 8 bits after band-limiting. The digital band-pass filter in the signal path is a four-pole Butterworth recursive filter. The correlators are implemented as full complex products, using TRW 8x8 bit multipliers. The product samples are accumulated during the correlation estimation period to reduce the correlation measurement flucuation noise.

Path length matching is very difficult to achieve. Yet, phase errors in the correlation loops will cause the beamformer weights to cycle, introducing a psuedo-doppler frequency offset to the beamformer output signal. To minimize this effect and to maximize the convergence accuracy of the adaptive loop, a calibration procedure is performed periodically to measure the loop phase shift for each element channel, so that phase correction can be applied during the normal adaptive mode. This procedure turns on one beamformer path at the time and a common signal introduced into the array channels before the element sample power dividers is used allow a correlation measurement for each of the signal correlation paths and the noise correlation paths. Phase information derived from this calibration measurement can then be used to rotate the correlation phasors for each element of the complex correlation vector during the adaptive mode.

Beam steering vector estimates can be obtained from the adaptive control loop in the converged mode. Convergence can be detected by comparison of the gradient norm-squared with a threshold after a specified time. In the converged mode, the optimum weight vector is related to the beamsteering vector by

$$R_{nn}\underline{w} = \underline{t} \tag{55}$$

where R_{nn} is the noise plus interference correlation matrix. But this is the noise channel correlation vector estimate's mean value. Thus, a Kalman filter is used to further enhance the estimation of the correlation vector's mean value when the adaptive loop reaches its steady-state. The resulting Kalman filter estimate of the beam steering vector is used to initialize the open-loop pointing program which controls the ground station's beamformer's between adaptive calibrations.

The discussion has focussed on the maximum SNR algorithm, so far. However, it is a simple step to now consider the case in which the direction of arrival of the signal is known. If the measured signal correlation vector is replaced by a computed beam steering vector, the results will be the same. Thus, basically the same approach could be used, with the appropriate beam steering vector, for the area coverage nulling antenna application.

TECHNOLOGY NEEDS

The payload capacity of the Shuttle orbiter presents exciting possibilities for future applications of adaptive arrays in communication satellites. Adaptive arrays have a better chance of being included in the payload. But weight and power are still major concerns, because of the competition by other high technology, such as on-board signal processing. Thus, the adaptive array hardware needs to be budgeted as a small fraction of the total; perhaps, tens of watts and one hundred pounds for a 5000 pound payload. Speed requirements depend on the application. Radiation requirements are severe. Reliability requirements are near .90 for ten years.

Weight and power requirements favor millimeter wave apertures and low power digital processing for adaptive array applications in which high resolution is demanded with extremely deep nulls. Autonomous on-board nulling demands that the digital components be able to survive a severe radiation environment with high reliability. Millimeter wave amplifiers and/or low-noise down-converters are needed to set the front end noise figures in the array element channels, so that power dividers and weight designs are not constrained by their impact on the G/T performance of the array. In military communication applications, computer speed with low power is important, since jamming interference directly affects channel availability. Computer requirements are in the 1-2 MOPS throughput range with fixed point arithmetic. Memory capacity to implement the serial type of gradient update algorithms are in the order of 16K bytes of memory.

Reliability of autonomous operation are present concern, due to the environment and the impossibility of repair, except by replacement. Fault tolerant computers may have a future role. For the time being, FEC memory, self-testing for fault diagnosis and isolation by a ground-based operator, and block redundancy are the norm.

The lucrative commercial applications of semiconductor VLSI components do not help the specifications of these components for military applications. Semiconductor manufacturers do not need such severe environmental specifications in commercial applications as in space and the military applications. Because of the high volume and less stringent requirements, the costs are low. There is a need for high reliability, low volume VHSIC and VLSI components which can satisfy the space environmental specifications. A price somewhat higher than commercial product levels might well be justified for the extra performance. Universally applicable standard functional modules would increase the volume of sales to offset the higher costs of production.

In the future, array antenna beamforming and adaptive processing will be all digital or, perhaps, optical. Digital sampling rates up to several hundred mega-samples per second with six to eight bit quantizing would be needed. Once the element channel signals were digitized, the beamforming and adaptive processing could all be performed digitally. The present bulky analog beamformers could be replaced by an inner-product chip or, more likely, all 20 beamformers in the present TDRSS system could be placed on one chip. Integrated optical systems could be used for beamforming and correlation operations. Because of the highly parallel processing optical throughput would be many orders of magnitude faster than computers would

could ever achieve. Input to the optical systems would be by means of LED arrays modulated with the array antenna signals.

THEORETICAL NEEDS

The major theoretical needs are in the algorithm area. There are several areas which need some development: performance criteria, performance prediction, and implementation. We have discussed the lack of a suitable performance criterion for specifying the phase of the beam steering vector, in the case when all of the users have equal EIRP and data rates and therefore presumably need the same information capacity through the multiple access channel. We have seen that experimental optimization of phase tilt in the nulling MBA can increase the resolution along one axis, as shown in the work of Potts, Mayhan, and Simmons (Ref. 3). But how does one formulate the goal that this process generates?

Performance prediction is most often obtained by computer simulation, at present. Prediction of the effect of weight tolerances is difficult, analytically. For example, it is known that if the signal and noise covariance matrices commute and the weights are subject to Gaussian distributed errors, then the SNR has a singly non-central F-distribution. However, the most interesting cases occur when there is a jammer signal present in the noise covariance matrix and the jammer is located at such a position that these matrices do not commute. The algorithms for maximum SNR are nonlinear, but interesting.

What we call theoretical needs for implementation sounds contradictory. However, the architecture of the adaptive antenna seems to have a processing bottleneck, at present. Direct matrix inversion has been shown to have great speed advantages over gradient algorithms when the weight updates are based on sample correlation estimates in both cases, rather than a filtered correlation estimate. However, the complexity of correlation measurement is a draw-back to the direct matrix inversion approach. There may be distributed computational architectures based on array or network computers in which only neighboring correlations would need to be measured. Or one might use redundant structure of the correlation matrix to advantage in factoring the matrix into sparse matrix factors, as in the FFT, for some array configurations.

There are plently of analysis problems in the current approaches, too. Optimum strategies for dealing with pulsed jammers need to be found. What is the best strategy to use against a partial band or frequency-following jammer: full-band adaptive processing, with its technological demand for high bandwidth and close tolerance components, or intermediate band processing after dehopping, with the need for fast convergence and look-ahead?

CONCLUSIONS

This paper was intended as a brief update and summary of work done over the last eight years in applying adaptive algorithms to communication satellites. The most significant product of this work is the MA system and its adaptive calibration system in the TDRSS[1]. The algorithm which was formulated to solve the maximum SNR problem, including the AGC gain and weight normalization algorithm equations seem well suited to the TDRSS application. Ground simulations of the TDRSS forward and return link through the MA ground system, beamformers and adaptive controller verify that the system can work within the specifications of 0.5 dB degradation from theoretical signal to noise improvement. Soon, TDRSS will be in position and we have an opportunity to verify the performance with an operational satellite system.

The analytical approaches presented here are differ in some details from the literature, because the performance criterion is different. But the rather large body of literature on adaptive algorithms provides some general principles which are applicable over a broad class of algorithms.

It seems that several performance criteria lead to the same optimum weight vector solution. This should not lessen the search for new and relevant performance criteria, because we are presently working with a narrow set of scenarios for adaptive behavior. Vector optimization is just beginning to be considered, for example, in which several aspects of performance are simultaneously optimized.

Algorithm research seems to be the main focus in adaptive systems. It is certainly the most interesting and intellectually challenging. The fact that there is a large base of support in related mathematical disciplines such as least-square estimation theory, control theory, the quantum mechanics, and numerical analysis should provide a strong momentum to the eclectic reader in this type of research.

Some technological and theoretical needs were discussed which, if satisfied, would promote the application of adaptive arrays to communication satellites. In these applications adaptive arrays offer the advantages of flexible reuse of antenna apertures by many different signals without significant performance degradation from theoretical, and the ability to automatically spatially filter the incoming signals to prevent interference. The feature of automatic interference rejection will become increasingly important in more congested frequency bands of the future as we become more dependent on satellite communication.

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[1]The MA system concept was first brought to this author's attention by NASA through reports of studies done by AIL (Ref. 15). The algorithm developed in TDRSS development studies by the author was generated independently of the AIL system, which was not fully revealed in these reports.

REFERENCES

1. DuPree, J.E., "TDRSS Multiple Access Receiving Phased Array System", ITC'78 Proceedings, pp. 243-249.

2. Weber, C.M., Haldeman, D., and Blyth, R., "TDRSS Telecommunications MA Return Channel", ITC'78 Proceedings, pp. 1107-1115.

3. Potts, B., Mayhan, J.T., and Simmons, A.J., "Some Factors Affecting Angular Resolution in an Adaptive Antenna", ICC'81, Paper 10.4.

4. Simon, H., "The Architecture of Complexity" in The Sciences of the Artificial, Cambridge, MA, The MIT Press, 1969.

5. Widrow, B., *et. al.* "Adaptive Antenna Systems", IEEE Proceedings, Dec. 1967, pp. 2143-2159.

6. Widrow, B., *et. al.* "Adaptive Noise Cancelling", Principles and Applications", IEEE Proceedings, December 1975. pp. 1692-1719.

7. Widrow, B., *et. al.*,"Stationary and Nonstationary Learning Characteristics of the LMS Adaptive Filter", IEEE Proceedings, August 1976. pp. 1151-1162.

8. Riegler, R.L., and Compton, R.T., Jr., "An Adaptive Array for Interference Rejection", IEEE Proceedings, June, 1973. pp. 748-758.

9. Howells, P.W., "Intermediate Frequency Side-Lobe Canceller," U.S. Patent 3202990, August 24, 1965.

10. Applebaum, S.P., "Adaptive Arrays", Syracuse University Research Corp., Report SPL TR66-1, August, 1966.

11. Griffiths, L.J., "A Simple Adaptive Algorithm for Real-Time Processing in Antenna Arrays," IEEE Proceedings. pp. 1696-1704.

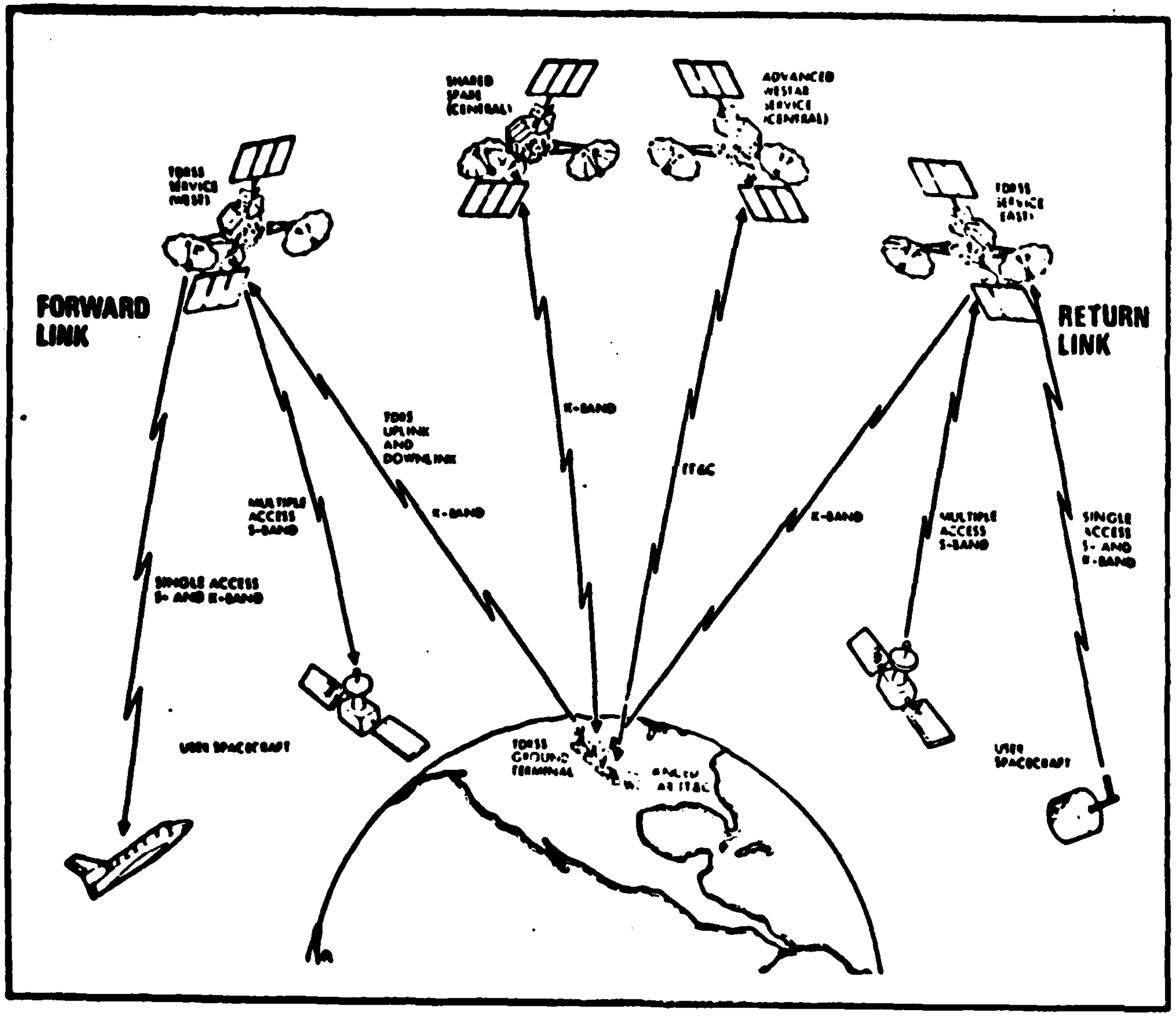
12. Balakrishnan, A.V., "Basic Concepts of Information Theory". Chap 5 of Information Theory, New York, McGraw-Hill Book Co., 1968.

13. Levine, R. D., and Tribus, M., The Maximum Entropy Formalism, Cambridge, MA, The MIT Press, 1979.

14. Widrow, B., "Stationary and Nonstationary Learning Characteristics of the LMS Adaptive Filter", IEEE Proceedings, August 1976. pp. 1151-1161.

15. AIL, a division of Cutler-Hammer, "Adaptive Ground Implemented Phased Array," Final Report, Prepared for NASA/GSFC, Contract NAS5-21653, Feb. 1973.

FIGURE 1. TDRSS



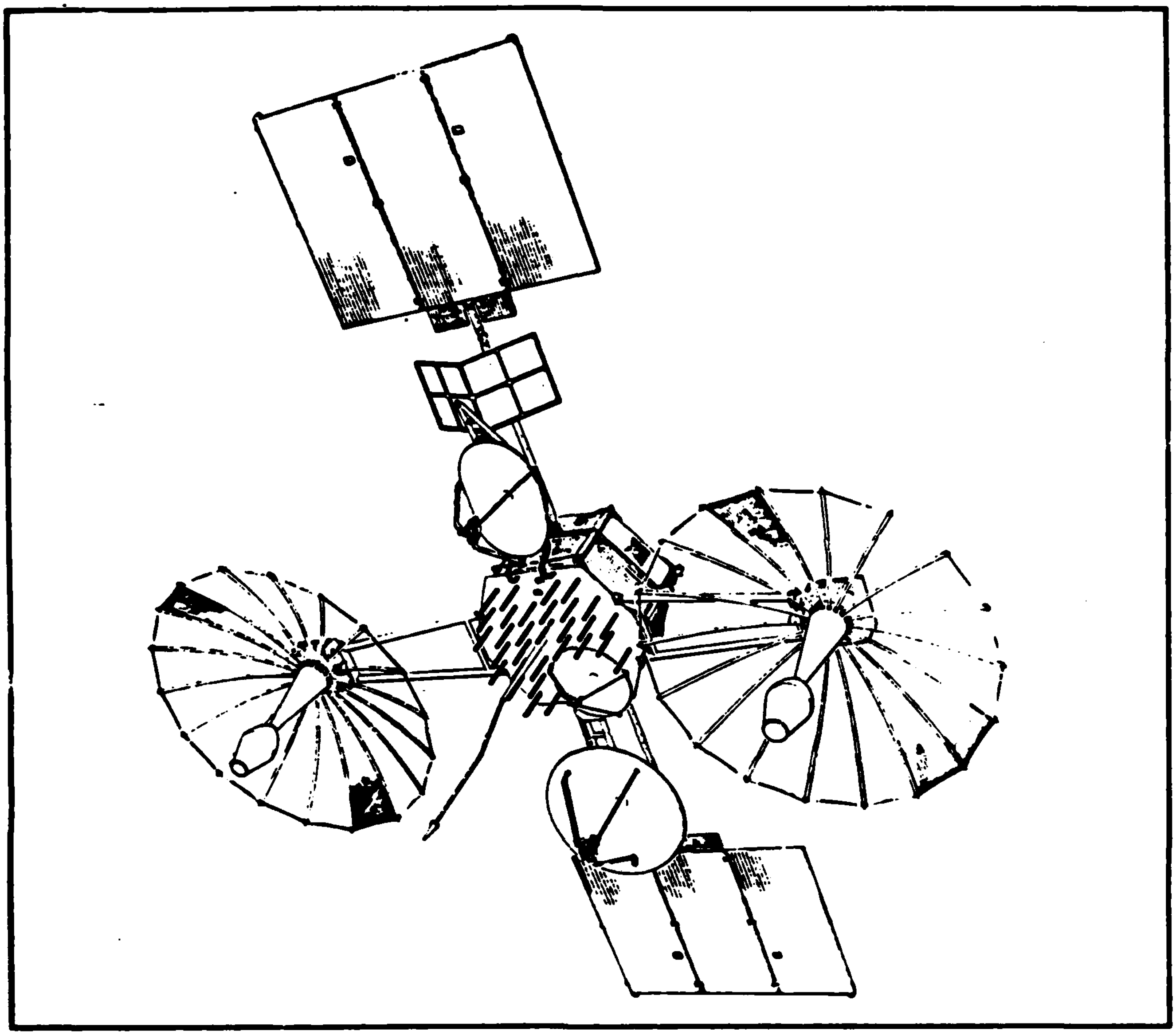


FIGURE 2A. SPACECRAFT CONFIGURATION

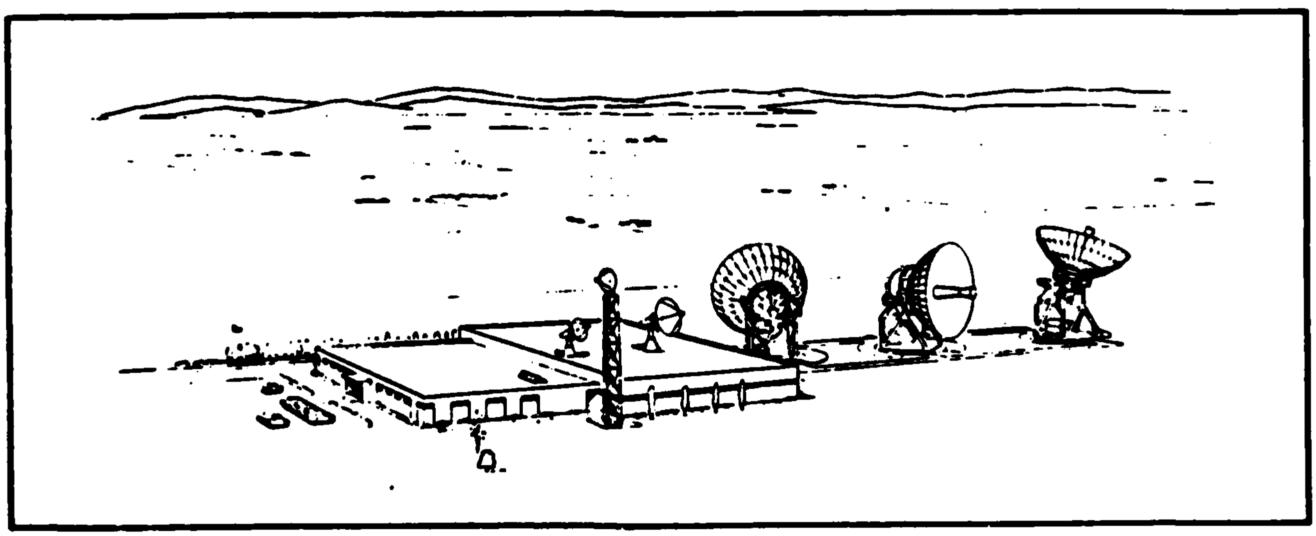


FIGURE 2B. GROUND SEGMENT ARTISTS CONCEPTION

FIGURE 3.

TDRSS MULTIPLE ACCESS RETURN LINK

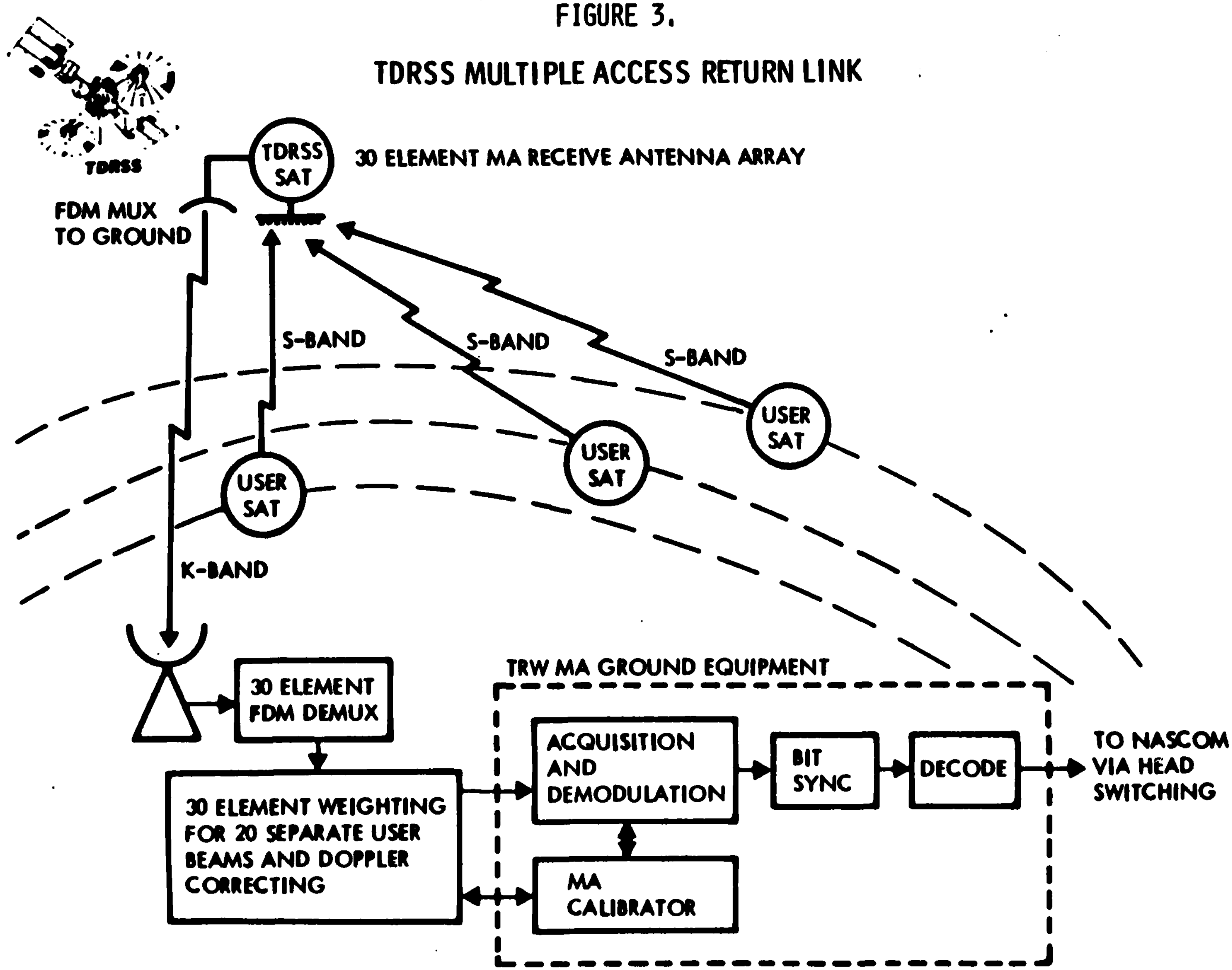


FIGURE 4.

CONTROL AND CALIBRATION CONCEPT

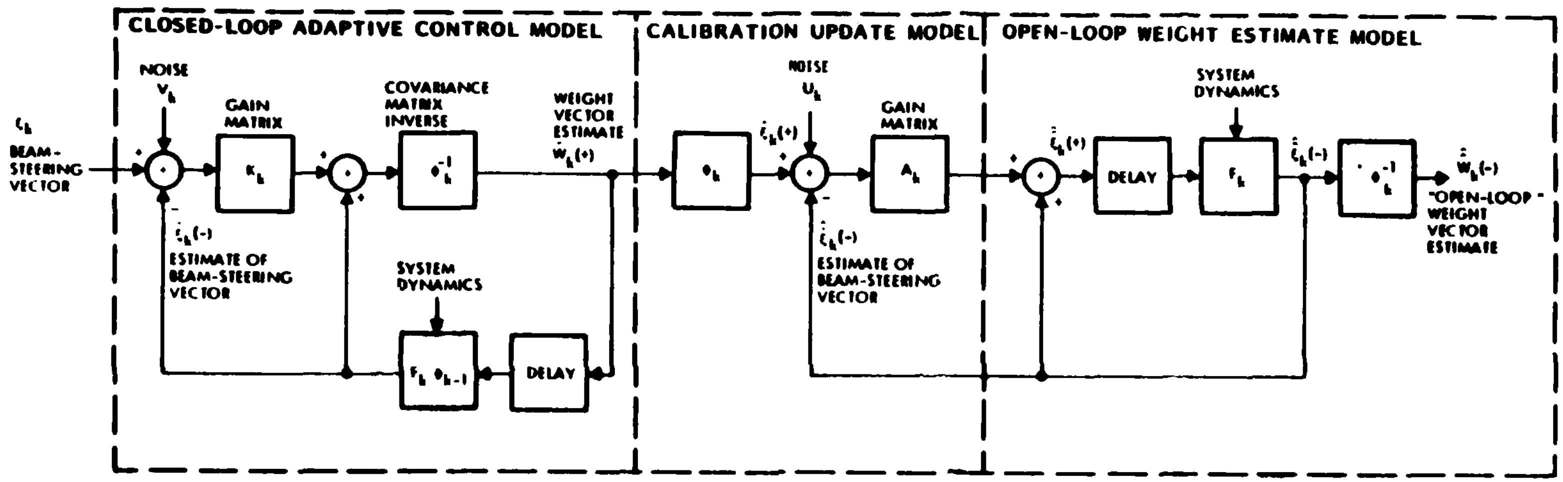
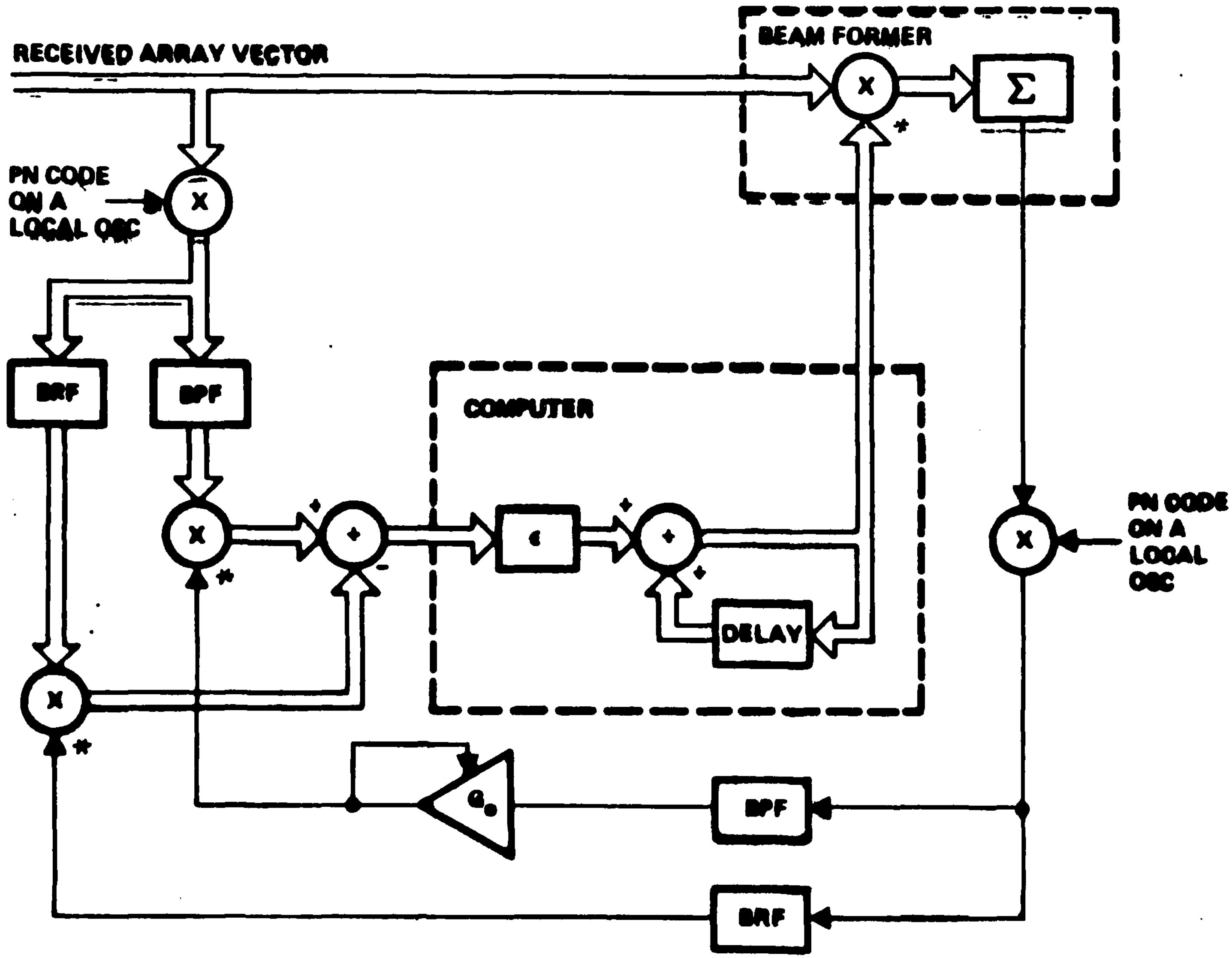


FIGURE 5.

MAXIMUM S/N ARRAY PROCESSOR



ENHANCEMENT OF SPREAD-SPECTRUM SYSTEMS THROUGH ADAPTIVE ANTENNA TECHNIQUES*

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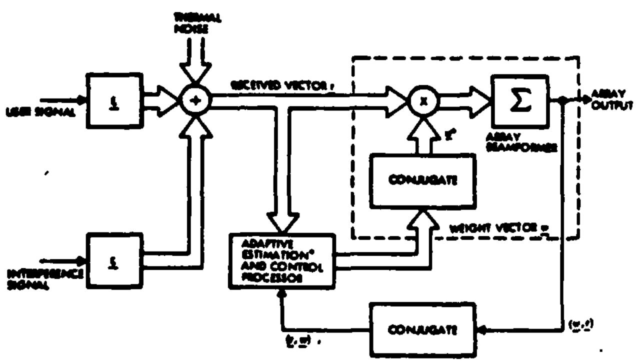
Abstract

Although spread-spectrum modems which exploit the structure of the desired signals in noise are highly developed, adaptive antenna techniques can further enhance the spread-spectrum performance by making the most effective use of the spatial properties of the signal and spatially distributed noise. This paper discusses the relative S/N advantage of optimum adaptive spatial filtering, which forms nulls, over simpler antennas which form beams in the user directions and depend upon low sidelobes for interference rejection. This relative advantage is expressed as an enhancement ratio that depends upon the interference signal parameters and upon the beam correlation, which relates the geometric and array-structure parameters. These results show when it is useful to add adaptive nulling to the spread-spectrum system, given an array structure and interference environment. Also, the multiple access use of optimum adaptive arrays is discussed and bounds on achievable S/N pairs and an optimum allocation procedure are presented.

Introduction

In recent years, either highly developed spread-spectrum or adaptive array systems were used to enhance communication system performance in an interference environment. Combined systems, incorporating both an adaptive array and a spread-spectrum modem, provide further enhancement over the interference rejection of either spectral or spatial filtering techniques alone. This paper discusses the performance limits of the integrated spread-spectrum modem/adaptive array processor.

An adaptive array is an array of antenna elements together with an adaptive processor. (See Figure 1.) The adaptive processor estimates signal and noise (including interference) parameters and uses an algorithm to optimize a set of complex weights according to a selected criterion such as minimum mean square error or maximum signal to noise ratio. Extensive discussions of adaptive array algorithms exist in the literature already. The brief sampling of algorithms presented in this section is to orient the readers who are new to this field.



*Incorporating spread-spectrum processor for signal and interference estimation

Figure 1. Adaptive Spatial Filter Concept

Many adaptive array processors use some version of Widrow's LMS (Least Mean Square) algorithm,[1] a gradient method of adaptively solving the normal (or matrix Wiener-Hopf) equations of the least-squares approximation problem. The LMS algorithm applies the orthogonal projection theorem of estimation theory[2] by adjusting the complex array weights to make the error between the array output and a reference signal orthogonal to the received element signals.

Applebaum's criterion,[3] maximizing the array output signal/noise ratio, is mathematically equivalent to a constrained optimization of a Hermitian form. The optimum complex weight vector satisfies a generalized eigenvalue problem. In the multiple user case, the analytical solution becomes quite involved. However, a closed form solution of this eigenvalue problem is quite easily derived in the single access case. The optimization problem can be stated in such a way that the closed form solution is identical to the solution of the LMS problem.

Since a closed form solution for the single access case is known, it is possible to compute it directly. Reed, Mallet, and Brennan,[4] showed that direct inversion of the estimated sample covariance matrix, though requiring high circuit complexity, leads to a rapidly convergent optimal processor for a small number of elements. Miller[5] compared the transient performance of six different algorithms and demonstrated that the direct approach gives more rapid convergence than the LMS algorithm. However, the number of calculations required per unit time is approximately proportional to m^3 and the number of circuit components is approximately proportional to m^2 where m is the number of complex weight controls.

Accelerated convergence and reasonable complexity is obtained with an intermediate approach: the recursive estimators. Baird[6] showed that the LMS adaptive processing structure is derived in a more general form by using the Wiener-Hopf equation as the observation equation in a Kalman filter[7] formulation. A priori knowledge is used to secure more rapid convergence, and dynamical models of the optimal array weights permit more accurate tracking in the dynamic environment of a rapidly moving platform, for example. This approach provides a more general implementation of the maximum S/N algorithm by using the necessary condition as the observation equation in the Kalman algorithm.

These references and their associated bibliographies establish the existence of solutions to the single access adaptive array problem. Questions of implementation in a given application, convergence speed and stability, and the effect of tolerances are each worthy of a complete paper. However, other questions of a (broader) theoretical nature are addressed here. We assume that the adaptive algorithm computes an exact solution to the optimization problem and study the results. The first results are general expressions

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for the theoretical S/N ratio degradation for the single user, multiple-interferer, scenario before and after adaptation. A well-known matrix inversion lemma is applied to the type of covariance matrices encountered in the adaptive array problem. The maximum enhancement of the S/N performance occurs when the interferers are at the 3 dB points of the beam correlation. The maximum enhancement ratio depends upon the maximum achievable interference to noise-plus-interference for a given interferer, and is asymptotic to a value 6 dB less than this number.

In the multiple access case, two or more users share the same antenna array pattern, but orthogonal signal design (for example, FDMA) is used to prevent interference between users. This requires an optimum allocation in addition to the signal to noise ratio maximization. Our approach maximizes the <u>weighted average</u> signal to noise ratio with <u>weighting factors chosen</u> to realize the desired allocation. Although the optimum allocation of multibeam communications satellite resources was discussed by Shaft and Roberts,[8] they did not consider the effect of an interference environment or correlation of user beams. The final results in this paper present the exact bound on the maximum achievable S/N pairs for a two user case. The achievable S/N pairs, in general, belong to a convex region analogous to the achievable rate pairs for the Gaussian multiple-access channel discussed in information theory by Shannon,[9] Cover,[10] and van der Meulen.[11] However, this bound is believed to be an original result in the context of antenna optimization.

Fundamental Background

The array (field) pattern of the spatial filter (Figure 1) has the form (+ = *T)

$$F(\vec{k}) = (\underline{w}|\underline{\xi}) = \underline{w}^{+}\underline{\xi} = \sum_{i=1}^{n} w_i^{*}\xi_i \tag{1}$$

where $\underline{\xi}$ is a vector with elements

$$\xi_i = \alpha_i(\vec{k}) \exp(j\vec{k}\cdot\vec{\rho}_i) \tag{2}$$

and $\alpha_i(\vec{k})$ is an amplitude factor describing the gain of each element as a function of the propagation vector $\vec{k}$. The factor $\exp(j\vec{k}\cdot\vec{\rho}_i)$ describes the phase advance of a wave transmitted in the direction $\vec{k}$ when the i^{th} element's phase center is located at $\vec{\rho}_i$, relative to the origin of the antenna's coordinate system.

The array output signal power due to a source producing an average power S per element from the "beam direction", $\underline{\xi}$, is proportional to

$$S|F(\vec{k})|^2 = S|(\underline{w}|\underline{\xi})|^2 = S\,\underline{w}^{+}\underline{\xi}\underline{\xi}^{+}\underline{w} \tag{3}$$

The dyadic, or outer product, $\underline{\xi}\underline{\xi}^{+}$, is a Hermitian matrix, since it has the property that it is self-adjoint; i.e., $A^{+} = A$.

An interfering point source producing an average power per element, J, after spread-spectrum processing in the beam direction, $\underline{x}$, produces an array output noise power proportional to

$$J|(\underline{\xi}|\underline{w})|^2 = J\underline{w}^{+}\underline{x}\underline{x}^{+}\underline{w} \tag{4}$$

In addition, the array output thermal noise power is proportional to

$$\sigma^2\sum_{i=1}^{n}\frac{\sigma_i^2}{\sigma^2}|w_i|^2 = \underline{w}^{+}\begin{bmatrix}\sigma_1^2 & & & \\ & \sigma_2^2 & & \\ & & \ddots & \\ & & & \sigma_n^2\end{bmatrix}\underline{w} \tag{5}$$

where

$$\sigma^2 = \frac{1}{n}\sum_{i=1}^{n}\sigma_i^2 \tag{6}$$

is the average thermal noise-power per element. The total array output noise power is the sum of independent thermal noises and interference. Let the noise covariance matrix be

$$\sigma^2\Phi_N = \sigma^2\begin{bmatrix}\sigma_1^2/\sigma^2 & & & \\ & \sigma_2^2/\sigma^2 & & \\ & & \ddots & \\ & & & \sigma_n^2/\sigma^2\end{bmatrix} + \sum_{i=1}^{n}\frac{J_i}{\sigma^2}\underline{x}_i\underline{x}_i^{+} \tag{7}$$

The noise covariance matrix is not only Hermitian, but also positive definite. (Spatially inhomogeneous thermal noise can be considered as a collection of a large number of weak point source interferers.)

The output signal-to-noise power ratio

$$R = \frac{S\,(\underline{w}|\Phi_S\underline{w})}{\sigma^2(\underline{w}|\Phi_N\underline{w})} \tag{8}$$

reduces to

$$R = \frac{S\,(\underline{w}|\underline{\xi}\underline{\xi}^{+}\underline{w})}{\sigma^2(\underline{w}|\Phi_N\underline{w})} \tag{9}$$

in the single user case. The object of optimal filtering is to choose the weight vector, $\underline{w}$, to maximize the ratio, R. A closed-form solution for the maximizing weight vector can be found by applying the Schwarz inequality.[12] Since the matrix, Φ_N, is Hermitian and positive-definite, it has an inverse and a square root which is also Hermitian and positive-definite.[13] Let

$$\underline{y} = \Phi_N^{1/2}\,\underline{w} \tag{10}$$

and

$$\underline{w} = \Phi_N^{-1/2}\,\underline{y} \tag{11}$$

Then,

$$R = \frac{S|(\Phi_N^{-1/2}\,\underline{\xi}|\underline{y})|^2}{\sigma^2(\underline{y}|\underline{y})} \le \frac{S}{\sigma^2}\,(\underline{\xi}|\Phi_N^{-1}\,\underline{\xi}) \tag{12}$$

by Schwarz's inequality. The upper bound is achieved for $\underline{y} = \Phi_N^{-1/2}\underline{\xi}$. Thus, the optimum weight vector is

$$\underline{w}_{opt} = \Phi_N^{-1}\,\underline{\xi} \tag{13}$$

By another approach, we may view this optimization as a variational calculus problem of finding the stationary points of a Hermitian form subject to a quadratic constraint. The method of Lagrange multipliers is used to maximize the output signal power with a constant output noise power constraint. Maximize the objective function,

$$H = S(\underline{w}|\Phi_S\underline{w}) + \lambda(c - \sigma^2(\underline{w}|\Phi_N\underline{w})) \tag{14}$$

The gradient of the scalar function, H, with respect to $\underline{w}$ is (to within a constant)

$$\nabla_{\underline{w}}H = S\Phi_S\underline{w} - \lambda\sigma^2\Phi_N\underline{w} \tag{15}$$

Nulling of the gradient, $\nabla_{\underline{w}}H = 0$, is a necessary condition for a stationary point. A second necessary condition is $\partial H/\partial\lambda = 0$. Thus, at the maximum S/N, the weight vector, $\underline{w}$, and the Lagrange multiplier, λ, satisfy the eigenvalue problem

$$S\Phi_S\underline{w} = \lambda\sigma^2\Phi_N\underline{w}\ , \tag{16}$$

subject to the constraint,

$$c = \sigma^2(\underline{w}|\Phi_N\underline{w})\ , \tag{17}$$

which demands constant output noise power. As was previously shown, the eigenvalue problem for the single access case has the solution given in Equation (13) where

$$\lambda = \frac{S}{\sigma^2}\,(\underline{\xi}|\Phi_N^{-1}\,\underline{\xi}) \tag{18}$$

is the maximum achievable signal/noise power ratio.

The solutions of Equation (16) are said to be eigenvalues of Φ_S relative to Φ_N. In the single access case, it can be shown that the optimum weight vector corresponds to the only nonzero eigenvalue of this problem. Thus, a gradient solution of this eigenvalue problem, corresponding to any nonzero (positive) eigenvalue, corresponds to the dominant solution.

A gradient solution of the eigenvalue problem is computed recursively using the rule

$$\underline{w}_{k+1} = \underline{w}_k + K_k\left[\nabla_{\underline{w}}\,H\right]_{\underline{w}=\underline{w}_k} \tag{19}$$

where λ_k is chosen as the signal-to-noise power ratio at each $\underline{w}$, and the gain matrix K_k is chosen at the k^{th} iteration to provide rapid convergence or small residual error in the presence of measurement errors. Examples of methods discussed in the literature are steepest descents, relaxation, and conjugate directions.

Single Access Performance Bounds

Suppose that an optimum spatial filter has adapted to a given interference environment. The maximum achievable S/N ratio is given by Equation (18) where the noise covariance matrix, Φ_N, describes the given noise and interference environment. Now, suppose that an additional point-source interferer is added. By considering the resulting S/N degradation before and after adaptation to the new interference environment, we can judge whether adaptation to the new environment is worthwhile and also judge the effect of various interference types, such as on-off pulses.

The noise covariance matrix changes to

$$\Phi_N(+) = \Phi_N(-) + \frac{J}{\sigma^2}\,\underline{xx}^+ \tag{20}$$

by the addition of an interferer of J/σ^2 average interference/thermal noise power ratio per element in the despread bandwidth. The interferer direction is described by a "beam vector" $\underline{x}$. Thus, its contribution to the noise covariance matrix is the dyadic (outer product) $J/\sigma^2\ \underline{xx}^+$.

The array output signal/noise ratio, using the weight vector obtained <u>before adaptation</u> to the new environment, is

$$R = \frac{S}{\sigma^2}\,\frac{|(\underline{\xi}|\Phi_N^{-1}(-)\,\underline{\xi})|^2}{(\underline{\xi}|\Phi_N^{-1}(-)\underline{\xi}) + (J/\sigma^2)|(\underline{\xi}|\Phi_N^{-1}(-)\underline{x})|^2} \tag{21}$$

This relation can also be expressed as

$$R = \frac{S}{\sigma^2}\,(\underline{\xi}|\Phi_N^{-1}(-)\underline{\xi})\left[1 + \frac{J}{\sigma^2}\,(\underline{x}|\Phi_N^{-1}(-)\underline{x})|\rho_{x\xi}|^2\right]^{-1} \tag{22}$$

where $|\rho_{x\xi}|^2$ is <u>defined</u> as the beam cross-correlation

$$|\rho_{x\xi}|^2 \triangleq \frac{|(\underline{x}|\Phi_N^{-1}(-)\underline{\xi})|^2}{(\underline{\xi}|\Phi_N^{-1}(-)\underline{\xi})(\underline{x}|\Phi_N^{-1}(-)\underline{x})} \tag{23}$$

By this means, we can interpret the degradation effect of adding an interferer <u>before adaptation</u> as a degradation factor

$$\left[1 + \frac{J}{\sigma^2}\,(\underline{x}|\Phi_N^{-1}(-)\underline{x})|\rho_{x\xi}|^2\right]^{-1} \tag{24}$$

(shown in Figure 2) that depends upon the maximum achievable interference to interference-plus-noise ratio in the given environment after spread-spectrum processing and with a beam correlation factor $|\rho_{x\xi}|^2$, which describes the geometric factors of array aperture design and spatial separation of the point source emitters. If $|\rho_{x\xi}|^2 = 0$, the new interferer is orthogonal to the optimum beam and there is no degradation. (In other words, the interferer is in a null of the array pattern.) Even if this is not the case, when the optimum interference to interference-plus-noise ratio is much less than unity in the despread bandwidth, there is little degradation. But if the product of the beam correlation and the optimum interference to interference-plus-noise is much greater than unity, there is a significant degradation. It will be shown later that as more interferers are added, their individual degradation impact is reduced.

To see the effect of S/N degradation after adapting to the new environment, one must invert the new covariance matrix and compare the optimum S/N obtained by adaptation with that available before. This is easily done by means of a matrix inversion lemma.

<u>Matrix Inversion Lemma</u>: If $\Phi(-)$ <u>is a positive definite Hermitian matrix and if</u> $\Phi(+)$ <u>is related to</u> $\Phi(-)$ <u>by</u>

$$\Phi(+) = \Phi(-) + a\,\underline{x}\underline{x}^{+} \tag{25}$$

<u>then, the inverse of</u> $\Phi(+)$ <u>is given by</u>

$$\Phi^{-1}(+) = \left[I - \frac{a\Phi^{-1}(-)\,\underline{x}\underline{x}^{+}}{1 + a(\underline{x}|\Phi^{-1}(-)\underline{x})} \right] \Phi^{-1}(-) \tag{26}$$

<u>where a is a positive real constant</u>.

The proof of this lemma is easily obtained by verifying that

$$\Phi^{-1}(+)\Phi(+) = I\,, \tag{27}$$

the identity matrix.

It follows from the matrix inversion lemma, applied to Equation (20), that the maximum achievable S/N after adaptation is

$$\frac{S}{\sigma^2}(\underline{\xi}|\Phi_N^{-1}(+)\underline{\xi}) =$$

$$\frac{S}{\sigma}(\underline{\xi}|\Phi_N^{-1}(-)\underline{\xi}) \left[\frac{1 + \frac{J}{\sigma^2}(\underline{x}|\Phi_N^{-1}(-)\underline{x})(1 - |\rho_{x\xi}|^2)}{1 + \frac{J}{\sigma^2}(\underline{x}|\Phi_N^{-1}(-)\underline{x})} \right] \tag{28}$$

The S/N degradation factor depends upon the maximum achievable interferer to interferer-plus-noise ratio and the beam correlation in a different manner than for the nonadapted case. A plot of this relation is shown in Figure 3.

By repeated application of the matrix inversion lemma, a sequence of matrix inverses can be generated, corresponding to each of the interferers contained in the environment without requiring any further explicit matrix inversions. This is especially useful if the matrix $\Phi(-)$ is a diagonal matrix. Note that the denominator in the matrix inversion lemma (26) is a scalar >1, since $a\underline{x}^{+}\Phi^{-1}(-)\underline{x}$ is a positive definite Hermitian form.

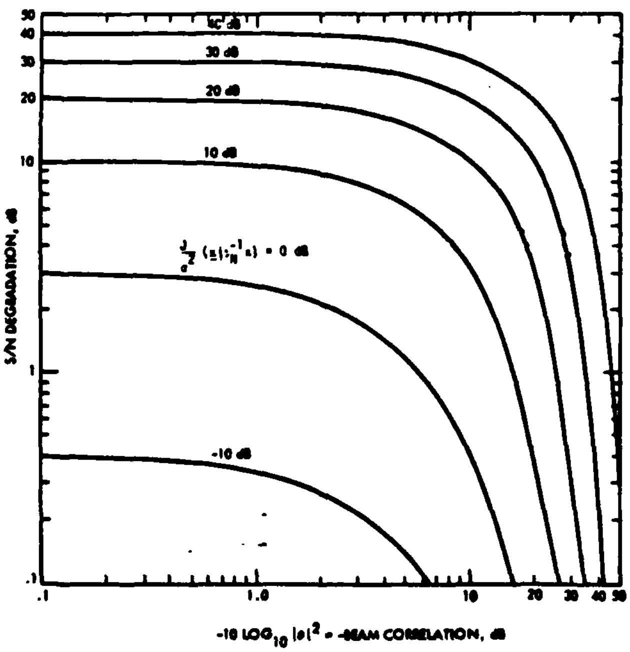


Figure 2. S/N Degradation Due to an Additional Interferer as a Function of Beam-Correlation and Maximum Interferer/Noise Ratio (Before Adaptation)

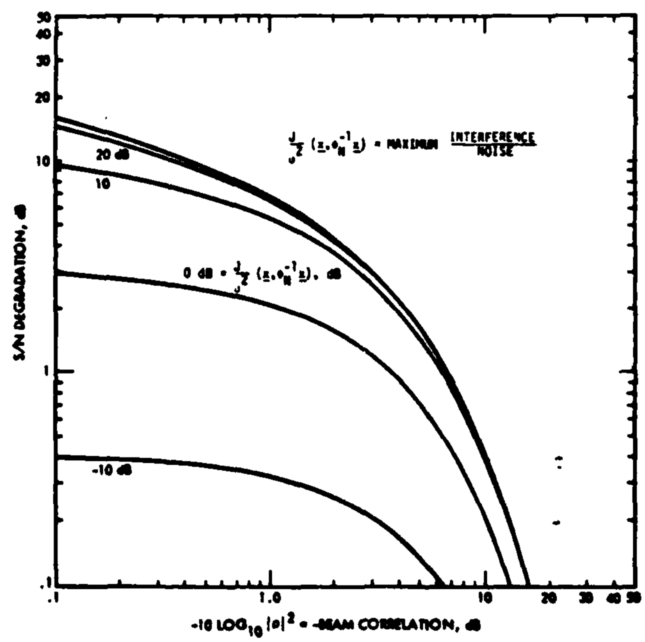


Figure 3. S/N Degradation Due to an Additional Interferer with Optimum Spatial Filtering (After Adaptation)

The enhancement ratio describes the ratio of the S/N degradation factors before and after adaptation to show the effect of optimum spatial filtering in improving the S/N ratio. The enhancement ratio, obtained from Equations (24) and (28) is (Figure 4)

$$E = \left[1 + \frac{J}{\sigma^2}\left(\underline{x}|\phi_N^{-1}(-)\underline{x}\right)\left(1 - |\rho_{x\xi}|^2\right)\right] \div \left[\frac{\left(1 + \frac{J}{\sigma^2}\left(\underline{x}|\phi_N^{-1}(-)\underline{x}\right)|\rho_{x\xi}|^2\right)^2}{1 + \frac{J}{\sigma^2}\left(\underline{x}|\phi_N^{-1}(-)\underline{x}\right)}\right] \tag{29}$$

Some special cases and examples explain the nature of this performance measure for the adaptive array. If the beams are orthogonal, $|\rho|^2 = 0$, $E = 1$, and the enhancement ratio is the same before and after adaptation (0 dB). If $|\rho|^2 = 1$, then $E = 1$, and the enhancement is again the same both before and after adaptation. In these special cases, adaptivity is useless. It is not needed in the first case since the interferer is already in a null. It is not possible to adjust the signal and interferer patterns independently in the second case since the beam correlation is 100 percent between the user and the interferer. Let us consider a more usual case in which enhancement is obtained.

Example 1. An interferer at -10 dB beam correlation has a maximum interference to noise ratio of, say, 10 dB. Before adaptation, Figure 2, the S/N degradation is 3.01 dB. After adaptation, the S/N degradation, Figure 3, is only 0.41 dB. The enhancement is S/N is 3.01 - 0.41 = 2.60 dB. This agrees also with the enhancement ratio curve shown in Figure 4.

In the special cases examined above, the unity enhancement ratio means that the adaptive system provides no enhancement at the extreme correlation values of 0 and 1. At intermediate values of beam correlation, as shown in Example 1, the enhancement ratio is greater than unity. Thus, the enhancement ratio must assume a maximum value in the beam correlation interval (0, 1). Indeed, the optimum value of beam correlation is $|\rho_{\underline{x}\xi}|^2 = 1/2$ and

$$E_{max} = \frac{\left(1 + \frac{1}{2}\frac{J}{\sigma^2}\left(\underline{x}|\phi_N^{-1}(-)\underline{x}\right)\right)^2}{\left(1 + \frac{J}{\sigma^2}\left(\underline{x}|\phi_N^{-1}(-)\underline{x}\right)\right)} \tag{30}$$

This maximum value of enhancement is asymptotic to a value which is 6 dB less than the maximum achievable interferer/interferer-plus-noise ratio in the despread bandwidth. That is, for

$$\frac{J}{\sigma^2}\left(\underline{x}|\phi_N^{-1}(-)\underline{x}\right) \gg 1 \tag{31}$$

then

$$E_{max} \cong \frac{1}{4}\frac{J}{\sigma^2}\left(\underline{x}|\phi_N^{-1}(-)\underline{x}\right) \tag{32}$$

A plot of E_{max}, Equation (30) is shown in Figure 5.

Example 2. An interferer of +20 dB J/N at -3.01 dB beam correlation causes 17.08 dB degradation before adaptation (Figure 3). After adaptation, the S/N degradation is only 2.97 dB. The enhancement ratio is 17.08 - 2.97 = 14.11 dB. This is the maximum achievable enhancement for a 20 dB interferer, as shown in Figures 4 and 5.

Multiple Access

The multiple access problem for the array is a maximization of the weighted average signal/noise ratio or, equivalently, a minimization of the array output noise power subject to an equality constraint on the

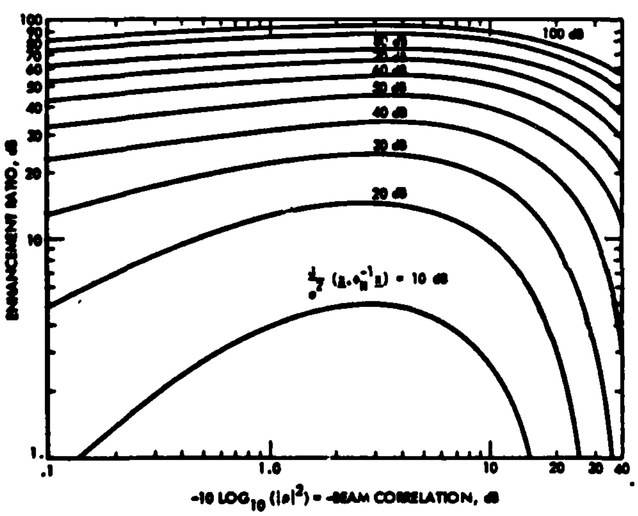


Figure 4. Enhancement Ratio = Relative Advantage of Optimum Filtering as a Function of Beam Correlation and the Maximum Interference-to-Noise Ratio

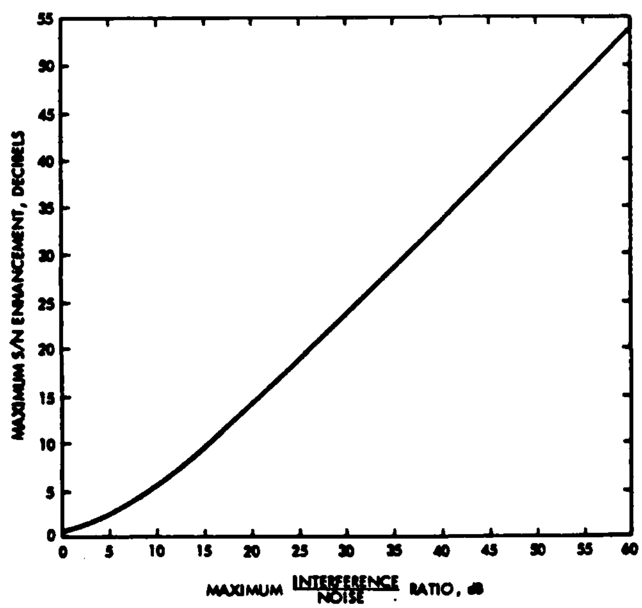


Figure 5. Maximum S/N Enhancement Ratio for Optimum Adaptive System

weighted average signal powers for each of the users. Thus, the optimum weight vector, $\underline{w}_{opt}$, must satisfy the eigenvalue problem

$$\phi_N \underline{w} = \lambda \sum_{i=1}^{m} \lambda_i S_i \xi_i \xi_i^+ \underline{w} \tag{33}$$

where we have assumed $\sigma^2 = 1$ without loss of generality and where the allocation weighting factors, λ_i, are chosen to satisfy the desired allocation. The form of the optimum solution,

$$\underline{w} = \sum_{i=1}^{m} \alpha_i \phi_N^{-1} \xi_i \ , \tag{34}$$

is the linear combination of solutions to the individual user maximum S/N problems.

Let the vector of weight vector projections in each user direction be denoted by $\underline{c}$ where $c_i = (\xi_i | \underline{w})$ is the i^{th} element of $\underline{c}$. Then, from Equation (34)

$$\lambda^{-1} \underline{c} = \begin{bmatrix} \lambda_1 S_1 (\xi_1 | \phi_N^{-1} \xi_1) & \lambda_2 S_2 (\xi_1 | \phi_N^{-1} \xi_2) \\ \lambda_1 S_1 (\xi_2 | \phi_N^{-1} \xi_1) & \lambda_2 S_2 (\xi_2 | \phi_N^{-1} \xi_2) \end{bmatrix} \underline{c} \tag{35}$$

for the case, $m = 2$, or

$$\lambda^{-1} \underline{c} = G(\lambda_1, \lambda_2)\, \underline{c} \tag{36}$$

where

$$\underline{c} = \begin{bmatrix} (\xi_1 | \underline{w}) \\ (\xi_2 | \underline{w}) \end{bmatrix} \tag{37}$$

and $G(\lambda_1, \lambda_2)$ is the Gram matrix in Equation (35). The eigenvalues are roots of the determinantal equation

$$\det[\lambda^{-1} I - G] = 0 \tag{38}$$

When $m = 2$, the eigenvalues λ^{-1} are roots of the quadratic equation

$$(\lambda^{-1} - \lambda_1 S_1 (\xi_1 | \phi_N^{-1} \xi_1))(\lambda^{-1} - \lambda_2 S_2 (\xi_2 | \phi_N^{-1} \xi_2)) - \lambda_1 \lambda_2 S_1 S_2 |(\xi_1 | \phi_N^{-1} \xi_2)|^2 = 0 \tag{39}$$

The dominant solution is

$$\lambda^{-1} = \lambda_2 S_2 (\xi_2 | \phi_N^{-1} \xi_2) \beta \tag{40}$$

where

$$\beta = \frac{1}{2}\left(1 + \alpha + \sqrt{(1-\alpha)^2 + 4\alpha|\rho|^2}\right) \tag{41}$$

$$\alpha = \frac{\lambda_1 S_1 (\xi_1 | \phi_N^{-1} \xi_1)}{\lambda_2 S_2 (\xi_2 | \phi_N^{-1} \xi_2)} \tag{42}$$

and

$$|\rho_{\xi_1 \xi_2}|^2 = \frac{|(\xi_1 | \phi_N^{-1} \xi_2)|^2}{(\xi_1 | \phi_N^{-1} \xi_1)(\xi_2 | \phi_N^{-1} \xi_2)} \tag{43}$$

is the beam correlation coefficient. The vectors ξ_1 and ξ_2 are orthogonal with respect to ϕ_N^{-1} if $|\rho|^2 = 0$.

If the vectors ξ_1 and ξ_2 are orthogonal with respect to ϕ_N^{-1}, and the roots of the determinantal equation are distinct, either

$$\lambda^{-1} = \lambda_1 S_1 (\xi_1 | \phi_N^{-1} \xi_1) \tag{44}$$

or

$$\lambda^{-1} = \lambda_2 S_2 (\xi_2 | \phi_N^{-1} \xi_2) \tag{45}$$

is the dominant eigenvalue. The corresponding eigenvectors are

$$\underline{c} = \begin{bmatrix} 1 \\ 0 \end{bmatrix} \quad \text{or} \quad \underline{c} = \begin{bmatrix} 0 \\ 1 \end{bmatrix} \tag{46}$$

One of the users is selected and the other is nulled. This is a classical capture effect. The selection of one or the other of the users is controlled by the choice of λ_1 and λ_2 to make its eigenvalue dominant. Both users are served when

$$\lambda^{-1} = \lambda_1 S_1 (\xi_1 | \phi_N^{-1} \xi_1) = \lambda_2 S_2 (\xi_2 | \phi_N^{-1} \xi_2) \tag{47}$$

is a double root of the characteristic equation. The normalized S/N ratio for the first user is then

$$\frac{|(\xi_1 | \underline{w})|^2}{(\xi_1 | \phi_N^{-1} \xi_1)(\underline{w} | \phi_N \underline{w})} = \frac{(\xi_1 | \phi_N^{-1} \xi_1)}{(\xi_1 | \phi_N^{-1} \xi_1) + (\xi_2 | \phi_N^{-1} \xi_2)} \tag{48}$$

and the normalized S/N ratio for the second user is

$$\frac{|(\xi_2 | \underline{w})|^2}{(\xi_2 | \phi_N^{-1} \xi_2)(\underline{w} | \phi_N \underline{w})} = \frac{(\xi_2 | \phi_N^{-1} \xi_2)}{(\xi_1 | \phi_N^{-1} \xi_1) + (\xi_2 | \phi_N^{-1} \xi_2)} \tag{49}$$

In general, $|\rho|^2 \neq 0$, and the ratio of user signal/noise ratios is proportional to the ratio of the gains. Then,

$$\frac{\lambda_1 S_1 |(\underline{\xi}_1|\underline{w})|^2}{\lambda_2 S_2 |(\underline{\xi}_2|\underline{w})|^2} = \frac{\alpha|\rho|^2}{(\beta - \alpha)^2} \tag{50}$$

This ratio is shown as a function of α in Figure 6. The classical capture effect is seen as the ratio α is varied. An enhancement occurs in the ratio of signal powers in the array output. When the correlation $|\rho|^2$ is small, the ratio of gains is a sensitive function α.

The bound on achievable S/N ratio pairs is determined as follows. Since

$$\lambda^{-1} \geq \frac{\lambda_1 S_1 |(\underline{\xi}_1|\underline{w})|^2}{(\underline{w}|\Phi_N \underline{w})} + \frac{\lambda_2 S_2 |(\underline{\xi}_2|\underline{w})|^2}{(\underline{w}|\Phi_N \underline{w})} \tag{51}$$

then, for the optimum weight vector, $\underline{w}$,

$$\lambda^{-1} = \left[1 + \frac{\lambda_1 S_1 |(\underline{\xi}_1|\underline{w})|^2}{\lambda_2 S_2 |(\underline{\xi}_2|\underline{w})|^2}\right] \frac{\lambda_2 S_2 |(\underline{\xi}_2|\underline{w})|^2}{(\underline{w}|\Phi_N \underline{w})} \tag{52}$$

and by substituting Equations (40) and (50) we obtain the equations for the maximum achievable S/N ratios:

$$\frac{|(\underline{\xi}_2|\underline{w})|^2}{(\underline{\xi}_2|\Phi_N^{-1}\underline{\xi}_2)(\underline{w}|\Phi_N \underline{w})} = \left[1 + \frac{\alpha|\rho|^2}{(\beta - \alpha)^2}\right]^{-1} \beta \tag{53}$$

and

$$\frac{|(\underline{\xi}_1|\underline{w})|^2}{(\underline{\xi}_1|\Phi_N^{-1}\underline{\xi}_1)(\underline{w}|\Phi_N \underline{w})} = \frac{\beta|\rho|^2}{(\beta - \alpha)^2 + \alpha|\rho|^2} \tag{54}$$

A plot of the maximum achievable S/N ratio pairs is shown in Figure 7 as α is varied, with the beam correlation as a parameter. This shows that two users cannot simultaneously realize their maximum achievable single access S/N ratios unless the beam correlation is unity. This means that two users are either collocated or located on periodic grating lobes of the array pattern. On the other hand, two users located within the half-power beamwidth of the array pattern can realize at least 86 percent of their maximum S/N ratios by properly weighting their signal powers in the average S/N.

Optimum allocation of the antenna gain in an interference environment is performed in the following way. The maximum S/N improvement for either of the users is equal to their improvement factors $(\underline{\xi}_1|\Phi_N^{-1}\underline{\xi}_1)$ or $(\underline{\xi}_2|\Phi_N^{-1}\underline{\xi}_2)$, respectively. If the gain is shared, then the improvement factor of each must be reduced. For given data rates, the user's threshold S/N requirements are determined by their signal design. *Allocate the gain in such a way that the two users reach their thresholds simultaneously as their signal powers are reduced by the same factor, in a given interference environment.* If the users do not threshold together, there is an imbalance in the allocation of antenna gain. Thus, set

$$\frac{S_1 |(\underline{\xi}_1|\underline{w})|^2}{S_2 |(\underline{\xi}_2|\underline{w})|^2} = \frac{T_1}{T_2} \tag{55}$$

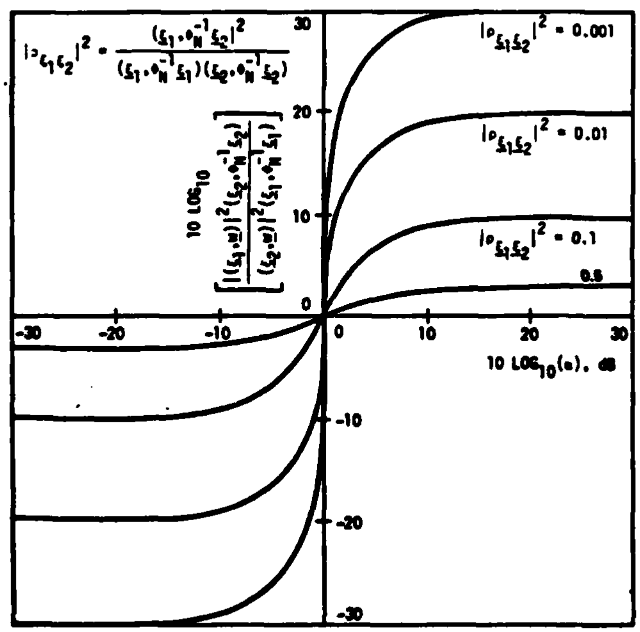


Figure 6. Normalized Gain Ratio Allocation Versus α

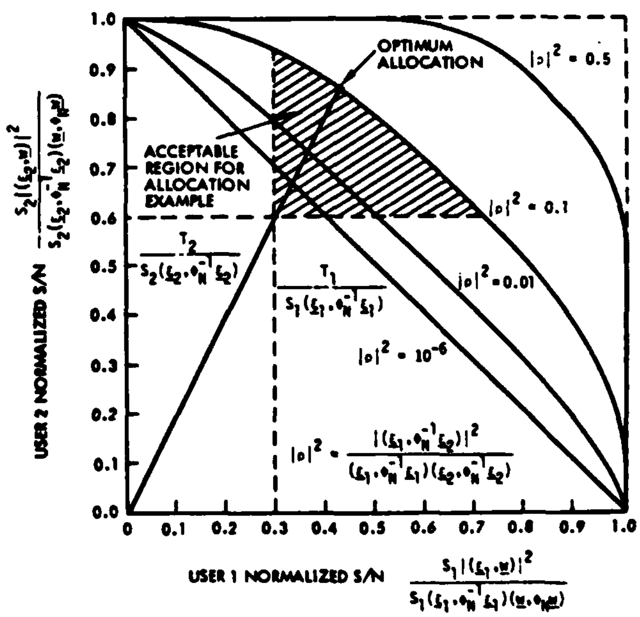


Figure 7. Achievable S/N Pairs for Two Multiple Access Users

where T_1/T_2 is the ratio of thresholds for the users. Then, in order to keep both users above their thresholds, we must have

$$\lambda^{-1} \geq \lambda_1 T_1 + \lambda_2 T_2 \tag{56}$$

The allocation problem is solved by finding the ratio λ_1/λ_2 from

$$\frac{T_1}{T_2} = \frac{\lambda_2}{\lambda_1} \frac{\alpha|\rho|^2}{(\beta - \alpha)^2} \tag{57}$$

Since we know T_1/T_2, and since both β and α are functions of λ_1/λ_2, we can solve Equation (57) for α, then determine λ_1/λ_2. The solution is

$$\frac{\lambda_1}{\lambda_2} = \frac{S_2(\underline{\xi}_2|\Phi_N^{-1}\underline{\xi}_2)}{S_1(\underline{\xi}_1|\Phi_N^{-1}\underline{\xi}_1)} \left[\frac{1 - |\rho| \sqrt{T_2'/T_1'}}{1 - |\rho| \sqrt{T_1'/T_2'}}\right] \tag{58}$$

for

$$|\rho|^2 \leq \frac{T_1'}{T_2'} \leq \frac{1}{|\rho|^2}$$

where T_i' is the normalized threshold of the i^{th} user, $T_i/S_i(\underline{\xi}_i|\Phi_N^{-1}\underline{\xi}_i)$. Table 1 shows examples for two cases.

Table 1. Optimum Gain Allocation for Two Users in Interference ($|\rho|^2 = 0.1$)

T_2/T_1	$\frac{S_1(\underline{\xi}_1\|\Phi_N^{-1}\underline{\xi}_1)}{S_2(\underline{\xi}_2\|\Phi_N^{-1}\underline{\xi}_2)} = 1$	$\frac{S_1(\underline{\xi}_1\|\Phi_N^{-1}\underline{\xi}_1)}{S_2(\underline{\xi}_2\|\Phi_N^{-1}\underline{\xi}_2)} = 2$	$\frac{S_1\|(\underline{\xi}_1\|\underline{w})\|^2}{S_2\|(\underline{\xi}_2\|\underline{w})\|^2}$
	λ_1/λ_2	λ_1/λ_2	
0.5	1.4045	0.5000	2.0000
1.0	1.0000	0.3560	1.0000
2.0	0.7120	0.2183	0.5000
3.0	0.5533	0.1294	0.3333
4.0	0.4366	8.0594	0.2500

To illustrate the application of the optimum weighting factors to gain allocation in an interference environment, suppose that two users have a beam correlation $|\rho|^2 = 0.1$ in an interference environment. User No. 1 has a threshold at 30 percent of his maximum achievable single user S/N. User No. 2 has a threshold at 60 percent of his maximum achievable single user S/N. To satisfy these threshold requirements, the allocation of S/N pairs must lie in the shaded convex region bounded by the curve for $|\rho|^2 = 0.1$ in Figure 7. Assume that the maximum achievable single user S/N's are equal. Then, $T_2/T_1 = 2.0$ and, from Table 1, $\lambda_1/\lambda_2 = 0.7120$. The optimum allocation point lies at the intersection of the boundary ($|\rho|^2 = 0.1$) with a line from the origin which passes through the intersection of the normalized thresholds, as shown in Figure 7. The optimum allocation point on the bound for $|\rho|^2 = 0.1$ assigns user No. 1 43 percent and user No. 2 86 percent of the maximum achievable single user S/N's, respectively.

Conclusions

The enhancement ratio, Equation (29), shows the relative advantage of optimum spatial filtering over a simpler array with low sidelobes. The enhancement ratio is a function of the maximum achievable interference/noise ratio of a new interferer introduced into the environment, where the noise includes all previous interferers. The maximum enhancement occurs for an interferer with a beam correlation of -3.01 dB and, for large interference/noise ratios, the enhancement ratio is 6 dB less than the interference/noise ratio.

The absolute S/N degradation saturates for interference/noise ratios larger than about 20 dB, and depends only upon the beam correlation for larger interferers. This gives an absolute worst case limitation to the S/N degradation for the ideal adaptive array. The beam correlation has a simple interpretation in a single interferer environment. It is the sidelobe response of an array pattern in the interferer's direction when a beam is formed in the desired user direction. Thus, the beam correlation factor contains all of the geometric factors related to array structure and user interferer geometry.

In the multiple user case, optimization of the array means maximization of a weighted average signal/noise ratio. The weighting factors in this average satisfy an optimum allocation of antenna gain in the interference environment considering the beam correlation between the users. Exact bounds on the maximum achievable S/N pairs are stated for a two-user case. The resulting curve, which encloses the convex region of achievable S/N pairs, represents the set of maximum achievable S/N pairs under all possible weightings in the average. This curve is a function of the beam correlation between the users. In the case of users orthogonal with respect to the noise environment, a capture effect can occur unless the weighting produces a double root of the eigenvalue problem. For a unity correlation, the users can each simultaneously realize their maximum achievable single user S/N's. The optimum allocation procedure results in simultaneous thresholding of both users and leads to a simple graphical procedure for evaluating the maximum achievable S/N pairs under an optimum allocation, using the bounds on the achievable S/N pairs.

References

1 Widrow, B., P. E., Mantey, L. J. Griffiths, and B. B. Goode, "Adaptive Antenna Systems," Proc. IEEE, Vol. 55, pp. 2143-2159, Dec. 1967.

2 Luenberger, D. G., Optimization by Vector Space Methods, New York, John Wiley & Sons, Inc., 1969.

3 Applebaum, S. P., "Adaptive Arrays," Special Projects Laboratory Report SPL-TR-66-1, August 1966, Syracuse University Research Corporation, Syracuse, New York.

4 Reed, I. S., Mallet, J. D., and Brennan, L. E., "Rapid Convergence Rate in Adaptive Arrays," IEEE Transactions on Aerospace and Electronic Systems, Vol. AES-10, pp. 853-863. November 1974.

5 Miller, T., The Transient Response of Adaptive Arrays in TDMA Systems, Ph.D. Dissertation, Ohio State University, 1976.

6 Baird, Charles A. Jr., "Kalman-Type Processing for Adaptive Arrays," ICC, 1974, Paper 10G.

7 Kalman, R. E., New Methods and Results in Linear Production and Filtering Theory, RIAS Technical 61-1, Baltimore, Md., Research Institute for Advanced Studies, 1961.

8 Shaft, Paul D., and J. A., Roberts, "Optimum Allocation of Multi-beam Communications Satellite Resources," IEEE Transactions on Communications, Vol. COM-24, No. 10, October 1976, pp. 1195-1200.

9 Shannon, C. E., "Two-way Communications Channels," in Proc. 4th Berkeley Symp. Probability and Statistics, Vol. I, Berkeley, Calif.: Univ. California Press, 1961, pp. 611-644.

10 Cover, T. M., "Broadcast Channels," IEEE Transactions on Information Theory, Vol. IT-18, pp. 2-14, January 1972.

11 van der Meulen, E. C., "A Survey of Multi-way Channels in Information Theory: 1961-1976," IEEE Transactions on Information Theory, Vol. IT-23, No. 1, pp. 1-37, January 1977.

12 Luenberger, Op. Cit., p. 30.

13 Peace, M. C., III. Methods of Matrix Algebra, New York, Academic Press, 1965, pp. 153-154.

14 Householder, A. S., "Numerical Analysis," in The Mathematics of Physics and Chemistry, Ed. by Margenan, H. and Murphy, G. M., Princeton, N.J., D. Van Nostrand Co., Inc., 1964.

MARLIN RISTENBATT

I'd just like to talk about two things. One is the effort that we are pursuing at the University of Michigan which is a small project that ties in very well with what we've been hearing heretofore. Then I'd like to lay down some thoughts I have concerning what's going on in the communications world in adaptive arrays. The thoughts are not entirely technical but deal with the sorts of things that we would like to attempt to do.

My favorite way to think about combining the adaptive array with spread spectrum in communications is to say that it plays the same role as power increases do. The bottom line performance of an antijam or any kind of ECCM application is that if you don't have to communicate while being too close to the jammer, then you don't need the transmitted power to be unusually high. The way we've been trying to lick the jamming problems of the world are to: (1) use spread spectrum; (2) increase the jammer power and (3) increase the transmitter gain if we can. Consider the adaptive array gain. See Slide 1b The G_R is the gain in the direction of the intended transmitter versus that in the direction of the jammer. P_J is the transmitted power of the jammer that results in power J at the receiver. Then there is the required E_b/j_0 for the system which may include the coding gain. Anyways, adaptive arrays can make a complementary contribution to spectrum spreading and power increases, except when the signal and jammer are aligned If I can get 20 dB from an adaptive array contribution, which comes solely from signal processing, I don't have to increase power and I don't have to use more spectrum. Of course, all the ECCM communicators now are trying to efficiently combine adaptive arrays with spread spectrum, and avoid any pitfalls of this combination. A tactical mobile communicator requires that the adaptive array be "fully adaptive"; in contrast, in many radar applications including sidelobe cancellors, we know the direction of arrival of the signal. If we don't know anything about the direction of the signal, which is the kind of a tactile mobile communication situation which I've always been thinking about, then we need the fully adaptive kind of model. Therefore, we are talking Slide 2 about the weights being computed as a product of a matrix inversion and a steering vector, where the steering vector strategy that we use is straightforward but very effective. That is, we divide the total sector into apriori trial periods. i.e., we do parallel processing. It's been used in the Navy with sonar for a long time, and it works very well here. In combining the adaptive array with spread spectrum, one thing that you could do (back to Slide 1) is put spatial processing in front of a modem. See Figure 1. One hazard, however, is that we may in fact cancel the intended signal during the (waveform) sync process, and we have the question of which algorithm, waveform or spatial, adapts first. This has sometimes been called the "great race" problem.

Simply placing spatial processing ahead of waveform processing is the simplest thing one could try, and that's not very sophisticated A more sophisticated approach was presented by Compton (it's been mentioned by Irving Reed), and that is, to put the spread spectrum modem in the loop of an analog LMS kind of algorithm. Slide 3 Now the array is in effect, trying to null anything that is not correlated with the intended signal. In this case the intended signal can be private to the communicators, consisting of a pseudonoise signal where we know the PN sequence, i.e., the system knows the sequence so that we should get the sum (in dB) of both processing gains. The system in Slide 3 is closed loop and Compton has claimed that this system

synchronizes very well; however it is in the sync area that it is most difficult to determine how well it works and under what conditions it works. It's that very point that I want to emphasize again and again, i.e., sync is the driver in these situations and that's where we need to work.

I've brought Slide 4 for a quick *refresher that all* matched filters of a pseudonoise signal will produce at their output the autocorrelation of the pseudonoise signal, i.e., we have the pulse compression of matched filtering a PN signal. We use this fact along with the open loop, sample matrix inversion *algorithm to do the following.* See Slide 5. We are doing the matched filtering processing first. This is contrary to the traditional way of putting the adaptive array processing first. This, we claim, comes very close in performance to the system which either does the processing *simultaneously, i.e., the array and temporal,* or does the temporal first. So, I will first do the matched filter processing on the received samples. We use the 2TW (where W is the instantaneous bandwidth) independent chip samples in the PN word for matched filtering, and use at least 2N (or all) of the same samples to compute noise covariance estimates from that same sample. We do not get in new data. We do the pseudonoise matched filtering as the first processing, and use the apriori steering beam. For the latter, we hypothesize signals in certain directions, and we did some studies as to how many directions we need in order to not lose sensitivity. We claim, (and it's been challenged by one person but I think that we are right) about 3 to 4 apriori directions is all you need to make use of the apriori sectors, i.e., you don't need a great many different apriori sectors in the steering sense.

Let me remind you that the pulse compression does two things for us. We're doing the SMI (sample matrix inversion) computation on hypothesized signal (using a sliding window processing) and we do the covariance estimate on the independent samples that we know do not contain signal if in fact the current hypothesized signal position is correct. So we have a disjoint signal-to-noise and noise-only condition. We're doing the covariance estimate in a communications problem and hence we don't know where the signal is in time, nor which direction it is. So we don't have the sidelobe canceller information, if you will. We can separate the noise-only values out to get the covariance estimate,based on 2N up to 2TW samples; we apply the algorithm on a sliding window basis and have increased the SNR, due to the matched filtering, for the spatial processing. That's what we're using.

Slide 6 I would like to just *summarize what I think is the status in the* general area of combining adaptive arrays with spread spectrum. First, I'm convinced that we must distinguish between some fairly disjoint application areas. The mobile tactical is the one that I was concerned with above, and within that we are talking about VHF, UHF and microwave frequencies. For the satellite cases, the sidelobe cancellor model may be of interest, as well as in-beam nulling. The HF and tropo scatter adaptive array models may also differ from the models and techniques useful in mobile tactical.

Let me comment that combining arrays with spread spectrum is an interdisciplinary business: Communications theory, antenna theory, and signal processing. My background has been in communications theory and spread spectrum. This topic is difficult to work in because you have to learn enough about one of the other areas, or the other two, to make any progress. As another

observation, it seems that there are a number of existing systems that are using new techniques, but we as researchers do not have a very good ability to learn how those things are going. This is because performance of the ECCM systems is usually classified. In order to be effective, researchers must get the feedback of how things are going, which lets them vernier and hone in on the correct problems. I do know of at least one case where a simple combination (of adaptive arrays and spread spectrum) was tried, and I am hoping that Jerry Gobien will tell you something about that. They had a fairly simple combination of adaptive array followed by a PN modem.

The current status is that placing a modem in a closed loop was proposed by Compton (as noted) and we have been pursuing the open loop, (matched filter plus the sample matrix inversion) which takes advantage of the transversal filter, stores the samples, and lets you use the same samples for the 2N up to 2TW computations. Our open loop technique is compatible with apriori sectors and exploits higher processing speeds that should be available with VLSI. By that we mean, it's true that the speeds inside the receiver processor are yet higher than the spread spectrum bandwidth, and we can take advantage of that. (A closed loop technique does not take advantage of that.) We claim that there are fewer sync problems with this technique, and we are currently trying to prove that.

The invitation to this workshop asked us to consider "what are the needs?" My answer is outlined on <u>Slide 7</u>. I claim that there is probably a unique combination of spread spectrum, adaptive arrays and error coding for each of those application areas that we've talked about. That's what we need to find. The driving problem is sync, and I think the optimum combination during acquisition sync is an open problem in the communications world. The radar is a somewhat different problem (85% of what you read about adaptive arrays applies to the radar situation). The communication applications (other 15%) are growing, and will continue to grow in the future.

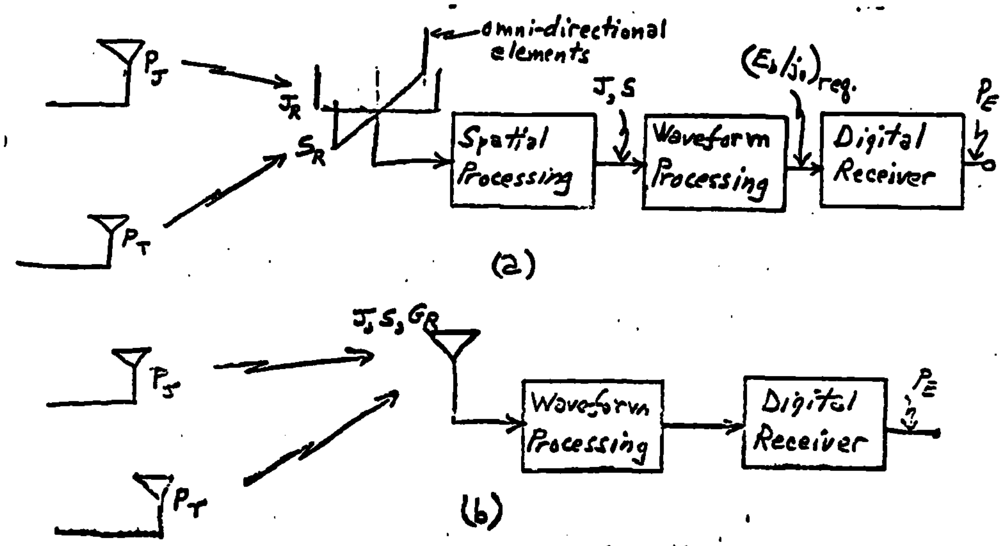
As noted, we must focus on the sync because that's the driver in the anti-jam world. I believe we must emphasize new architectures, such as the spectrum bus, of which JTIDS is an example. The conferencing net, which is now called Enhanced JTIDS (EJS) used to be called SEEKTALK (which is what Jerry Gobien is going to be talking about tomorrow). Also, there are self-organizing systems like packet radio. These are fairly new architectures and the way that adaptive arrays, spread spectrum and error coding are going to play together is still an open problem. We need network concepts to be developed concurrent with the technology pursuit, and we need better feedback from the user community.

<u>Slide 8</u> Finally, I want to talk about trends and opportunities. Open loop is the wave of the future. With VLSI, increased processing capability is coming. Transversal filters are the right way to go. The incoming samples are momentarily stored and you can use them for iterative calculations. I claim that two-way communication is getting to be integral with these new types of networks, i.e., the three that I've mentioned. When you have two-way communications, new solutions appear to some of the problems that we have had all along. For one thing we can envisage transmitter adaptive arrays. (All the adaptive arrays thus far have been receiver adaptive arrays.) Feedback may permit synergism for achieving ECCM that ought to help a great deal. (I'm talking tactical mobile where you don't know in advance the direction you need to go to.) In addition to transmitter arrays, it also permits power control. I fully believe that

if we had a satellite in the sky looking down at all of us, making judgments, they would say that our attempts to solve our ECCM problems solely by increasing power and taking more bandwidth may in some cases, be creating as many problems as it solves. We are interfering with ourselves and we are taking inordinate bandwidth. So power control comes with this two-way business. And finally, we can also have adaptive data rate so that we can use the system to the maximum rate that the jamming will prevent. Also we might improve and refine the new network architectures that have emerged. That concludes what I want to say.

Adaptive Antenna plus Spread Spectrum

Spatial Processing (Adaptive Antenna) — Waveform Processing (Spread Spectrum)



Receiver Block Diagrams: a) Combined processing; b) Equivalent of (a)

The Covariance Matrix: M

Objective: M describes, in systematic form, all of the information about the interference environment.

Elements of M $\qquad M = (m_{k\ell})$

where: $m_{k\ell} = E(m_k^* m_\ell)$ = expected value of the product of the noise received by the k^{th} and ℓ^{th} antenna elements.

Optimum Weights:

$$MW = \mu S^* \Rightarrow W = \mu M^{-1} S^*$$

where: M = covariance matrix of interference
μ = arbitrary nonzero complex constant
S^* = complex conjugate of the desired-signal matrix (from the array)
W = complex weight vector

Open Loop $\longrightarrow$ Matrix Inversion

Closed Loop $\longrightarrow$ M values are the asymptotic values

SLIDE 2

AD-A138 229 RESEARCH TRENDS IN MILITARY COMMUNICATIONS: PROCEEDINGS 2/4
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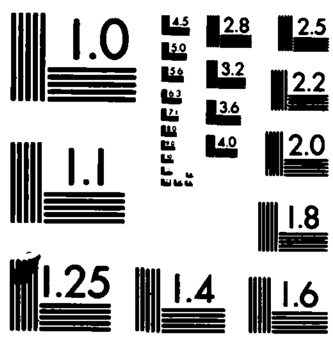
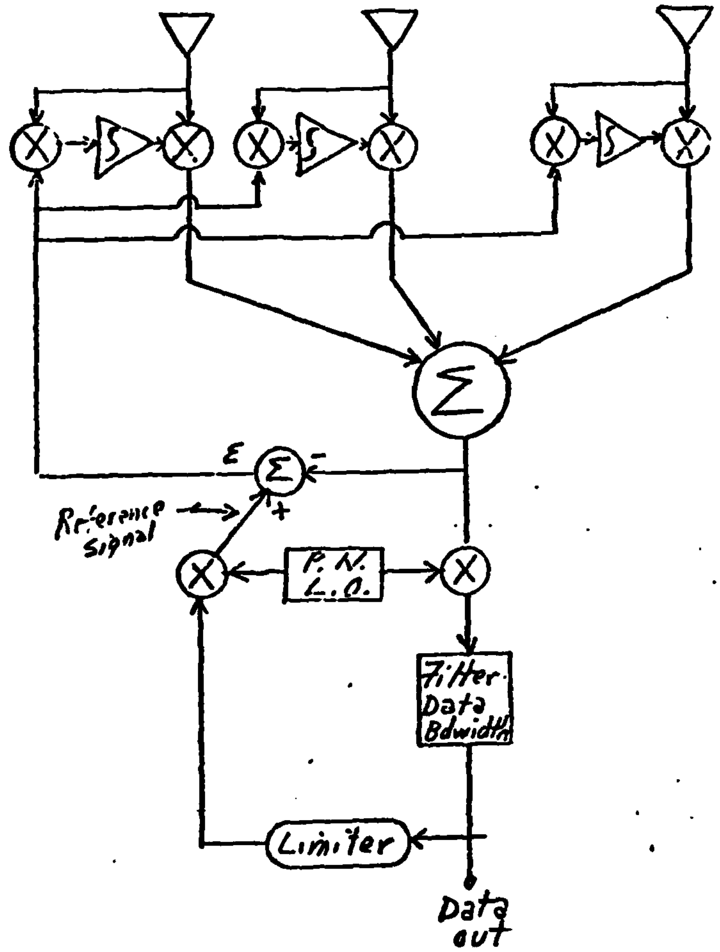


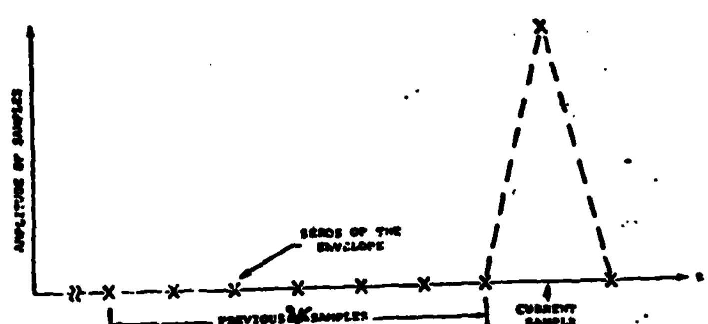
MICROCOPY RESOLUTION TEST CHART
NATIONAL BUREAU OF STANDARDS-1963-A

SLIDE 3



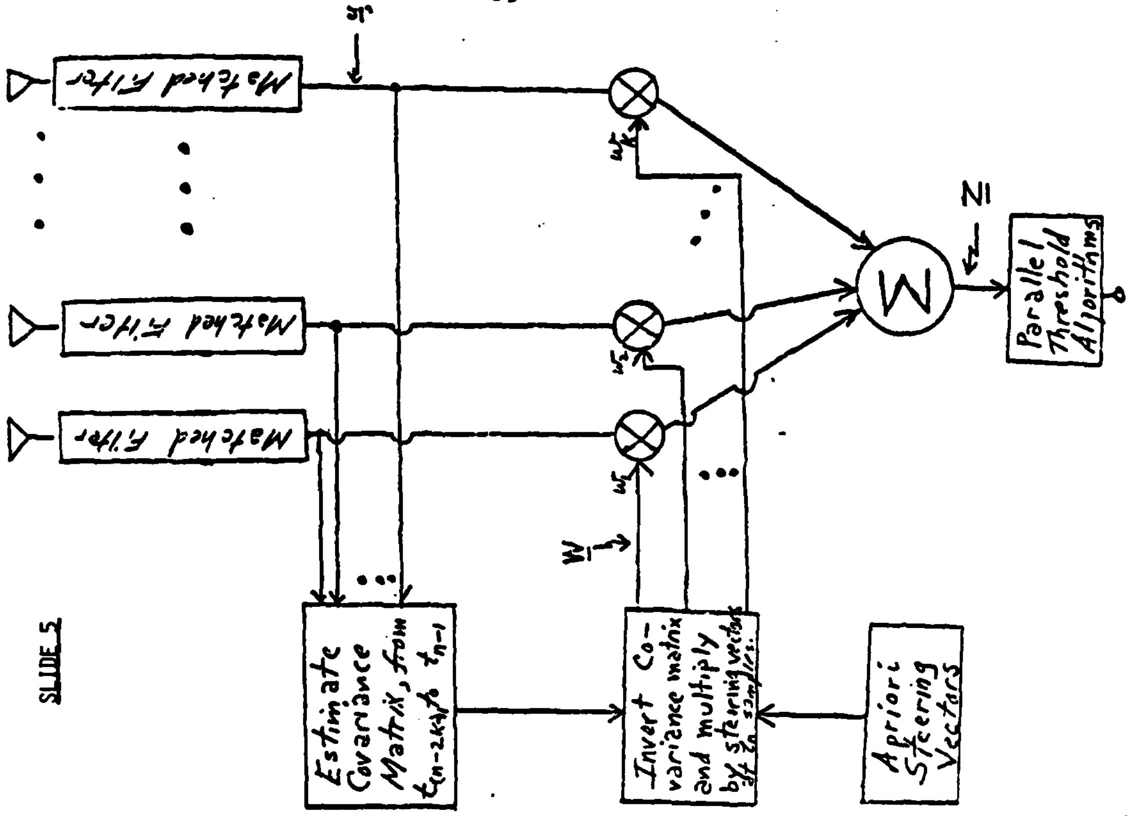
Closed-Loop Processor for PN, Using LMS Processor

SLIDE 4



Sampled Envelope Values of Signal Autocorrelation

SLIDE 5



- PN Matched Filters followed by SMI.

SLIDE 6

STATUS SUMMARY

APPLICATION AREAS
- Mobile Tactical Comm < VHF, UHF, μwave
- Satellite < UHF, μwave, EHF
- HF, Troposcatter

INTERDISIPLINARY < Comm. theory, Antennas and E&M, Signal Processing

EXISTING SYSTEMS — (Feedback limited)

SIMPLE COMBINATION < Closed loop, PN

MODEM IN CLOSED LOOP (Compton)

OPEN LOOP — M. Filter plus SMI
- Transversal filter
- A-priori Sectors — parallel processing
- Higher processing speeds
- Fewer Sync problems

SLIDE 7

NEEDS

UNIQUE COMBINATION OF
- SPREAD SPECTRUM
- Adaptive Array
- Error-Coding

FOCUS ON SYNC — FOR EACH APPLICATION AREA

EMPHASIZE NEW ARCHITECTURES:

SPECTRUM BUS (JTIDS)
CONFERENCING NET (HAVECLEAR)
SELF-ORGANIZING (Packet Radio)

NETWORK CONCEPTS — CONCURRENT WITH TECHNOLOGY PURSUIT

BETTER FEEDBACK FROM USER COMMUNITY

STRENGTHEN INCENTIVES/REWARDS — TO INCREASE PAYOFF

SLIDE 8

TRENDS/OPPORTUNITIES

OPEN LOOP
- OPEN ENDED
- VLSI

TRANSVERSAL FILTER EXPLOITATION

TWO-WAY COMM. IN NEW NETWORKS → Transmit ADpt. Array
POWER CONTROL
ADAPTIVE RATE

NEW NETWOK ARCHITECTURES

ROBERT DINGER

Slide 1 I'm Bob Dinger from China Lake. I'll talk about an array that I call the reactively steered adaptive array of RESAA concept [1,2].

Slide 2 I'll talk about the theory followed by our recent results on a 4 GHz microstrip array, both simulations and experimental measurements. I'll say up front that it's still an open question as to how exactly the RESAA approach is going to apply to a spread spectrum system I believe it could be used in a wide bandwidth system, and I think it does have promise for improved performance over other adaptive array techniques

Slide 3 Basically, the concept is quite simple. See Figure 1 You have one element in the array that s only connected to the receiver; the other elements are all parasitic. They are mounted in close proximity to get tight coupling, and the beam pattern is formed by the values of the adjustable reactive terminations. You monitor the the output of the receiver and use an adaptive algorithm to adjust the reactive terminations. Physically what is happening is that in addition to receiving the main signal, the center element is receiving scattered radiation from the parasitic elements. The scattered radiation phase depends very critically on the value of the reactive terminations

Slide 4 The theory looks deceptively simple. Roger Harrington [3] worked the theory out in 1978, and showed that for a parasitic array, the output is given by Eq. 1 on slide 4. The quantity $g_e\theta$ is the usual element pattern, along with the phase factor for each element. The quantity $[Z_A+Z_L]^{-1}$ corresponds to the driving current at each element and includes the mutual impedance matrix. In the load matrix, which contains the reactive terminations, the central element (p^{th}-element) connected to the receiver is matched Because the output goes as $[Z_A+Z_L]^{-1}$ this mixes together all the terminations so that the current in each element is a function of the setting of all the other elements. That also makes analysis hard to carry beyond this point.

Slide 5 Figure 2 is a plot of a 3-element array output as a function of the normalized reactive loads on the two parasitic elements for a signal incident broadside to the 3-element array You see that the choice of reactive loads that minimizes the output power of the array is given by the bottom of the bowl Therefore, any control algorithm basically estimates locally the surface gradient and attempts to go to the bottom of the bowl Unfortunately, it is not a parabolic bowl Of course the optimum point, (the solution point) moves around as a function of angle incidence and number of interference sources.

Slide 6 Now using Harrington's theory back at the Naval Research Laboratory (NRL) a few years ago we tried pattern adjustment deterministically with an HF ring array of 7 elements, where the center element was connected to the receiver [4,5]. See Figure 3. Triangles represent the pattern we are trying to synthesize by doing nonlinear least squares fitting on a computer. The fitting determined what values of reactive loads were needed. those values were set on the terminations, and then the pattern was measured. Figure 3 shows that we can do this fairly well except for the back lobes. Fig. 3 shows 3 trials. Harrington's original idea of doing this deterministically. is probably not practical. That is because you can't accurately set the reactive loads, and you don't know the theory of the mutual coupling and all its imperfections well enough. That's when we tried to do it adaptively.

Slide 7 Figure 4 shows a pattern for

an HF array at 23 MHz which is about 15 meters wavelength. The array was 80cm in diameter so that the overall size of the array is about 1/20th of a wavelength The interference incidence angle is indicated by the arrow. This response was obtained by manually adjusting the loads on the elements to minimize the received interference power The performance is quite good for an HF array that you can actually carry in your hands That work is continuing at NRL.

Slide 8 The array I want to present data on is a 4 GHz 5-element microstrip array Figure 5. The center element is connected to the receiver, and there are 4 elements with reactive loads. The loads are simple varactor phase shifters Figure 6. Figure 7 shows the kind of performance that we obtained last year again by manual adjustment of the terminations. The arrow is the interference signal and you can null it down to -30 dB or -40 dB below the *average value* of the pattern away from the null.

* **QUESTION**: What's the element separation in wavelengths ?

* **ANSWER**: The separation is a tenth of a wavelength between elements.

The rest of this talk concerns my recent result in automatic adaptive control. The algorithm uses the steepest descent equation which is shown on Slide 9. The details of how you estimate the gradient and exactly how the order of the steps are done for control are not critical to this discussion. Basically you have a rate constant K, which, if too high, produces noise. You will get faster conversion, but the jitter in the "bottom of the bowl" is too much. Possible criterion functions include, for power inversion with no reference, just the output of the array (V_o), or the average of that output If you can derive an estimate of the interference-to-signal ratio you could use that to adapt the array

Slide 10 Figure 8 gives some simulation results starting with initial normalized reactance values of 0.2 This shows how the weights adapt themselves to the final steady state solution. There are two examples of K shown here. One is a K that's perhaps a little too small; the other is a K that's too large. The sawtooth is jitter at the optimum solution.

Slide 11 Figure 9 shows the beam history for Figure 8. Iteration 0 shows the initial beam and it adapts for interference at -45 degrees, the beam forms a null This is a simulation for 3 elements.

Slide 12 Figure 10 shows a measurement system for an array that is actually 4 parasitic elements (there are only 2 shown). You have a test signal, the array is controlled by a computer, etc.

Slide 13 Figure 11 presents some results of interference reduction. The initial power is about -20 dBm. There are 2 different starting solutions, i.e., the voltage of the varactor diodes in volts shown The receiver power is reduced from about -20 dBm down to around -50 or -60 dBm The jagged curve is jitter at the optimum solution.

Slides 14, and 15 Figure 12 is the reactive load variation during the power variation shown in Figure 11. Figure 13 shows an example in which we track during antenna rotation. The system first converges, then tracks while the antenna rotates 45 degrees.

Slide 16 Figure 14 shows "snapshots" of the pattern during the antenna rotation. It shows how the null points towards the interference as it moves from 90 to 120 degrees.

Figure 15 shows an example of a

pattern formed with both a desired signal and a source of interference present The algorithm is based on the steepest descent method, with an estimate of the desired signal derived from a synchronous receiver. Further details of the receiver structure are given in [6] for this mode of operation.

In Figure 16 the nulling frequency response is given To obtain this plot, a null was formed broadside at a frequency of 4.00 GHz. The reactive load values for this null were then left fixed, and the frequency was swept. The nulling bandwidth, defined as the frequency range over which the interference is rejected by at least 20 dB relative to the pattern maximum, is about 20 MHz in this example. Figure 17 displays the patterns measured at selected frequencies in the range included in Figure 16 A change in frequency causes the null to change in angular width, depth, and azimuth angle.

The nulling frequency response is determined by the array geometry and the bandwidth of the reactive loads, the nulling bandwidth can expect to improve substantially with more attention paid to making the reactive load bandwidth as wide as possible (bandwidth has not been optimized for the single varactor diode design used in the array). The array bandwidth is about 100 MHz (about 2.5%), which is typical of microstrip patch elements. Operation over wide bandwidths, which would be necessary for a spread spectrum application, has not been a specific goal of the present study; however, such operation should be possible by using antenna elements and reactive loads with a wide bandwidth.

The advantage I see for this adaptive array technique is that a lot of hardware is eliminated, and possibly, a lot of hardware which limits the bandwidth. An hypothesis that I have is that you get better pattern control for an array of the same size; because you are controlling the coupling you can better control the pattern. I also think it has potential for an add-on array in front of an existing receiver, perhaps more so than for other adaptive array techniques.

REFERENCES

1. R.J. Dinger, "Adaptive Microstrip Antenna Array Using Reactively Terminated Parasitic Elements," 1982 AP-S International Symposium Digest, Albuquerque, NM, 24-28 May 1982, pp. 300-303

2. R.J. Dinger "A Microstrip Power Inversion Array Using Parasitic Elements," 1983 AP-S International Symposium Digest, Houston, TX, 23-26 May 1983, pp. 191-194.

3. R.F. Harrington, "Reactively Controlled Directive Arrays," IEEE Trans. Antennas Propag. AP-26, 390-397 (May 1978).

4. R.J. Dinger and W.D. Meyers, "A Compact Reactively Steered Antenna Array," 1980 AP-S International Symposium Digest, Quebec, Canada, 2-6, June 1980, pp. 312-315

5. R.J. Dinger and W.D. Meyers, "A Compact HF Antenna Array Using Reactively Terminated Parasitic Elements for Pattern Control," Naval Research Lab Memo Report 4797, May 1982.

6. R.J. Dinger, "Closed-Loop Adaptive Control of a 4.0 GHz Reactively Steered Microstrip Array: Experimental Results," Naval Weapons Center

Technical Publication 6451 (in press, available from author after October 1983).

ADAPTIVE ARRAY BEAMFORMING USING REACTIVELY-TERMINATED PARASITIC ELEMENTS

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SLIDE 1

5/83

OUTLINE

- Reactively-Steered Adaptive Array (RESAA) Concept
- Theory
- Early Work at HF (20 MHz) at Naval Research Lab
- 4.0 GHz Microstrip Array
    - * Simulations
    - * Experimental Results
- Applications to Spread Spectrum Communications

SLIDE 2

5/83

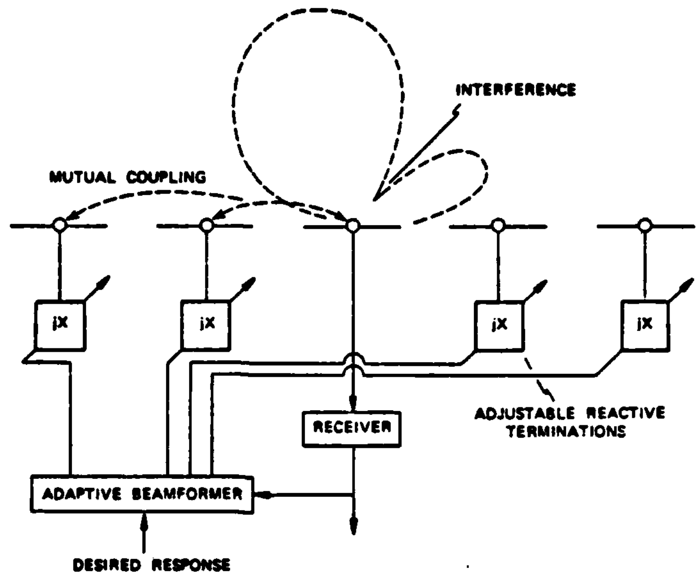


Figure 1.

SLIDE 3

THEORETICAL ANTENNA PATTERN

- As developed by Harrington (1978), receiver output is given by

$$V_o(\theta) = g_e(\theta) \sum_n \left\{ [Z_A + Z_L]^{-1} \right\}_{np} e^{jk_o x_n \cos\theta} \qquad Eq\ (1)$$

where

$[Z_A]$ = array mutual impedance matrix

$$[Z_L] = \text{load impedance matrix} = j \begin{array}{r} \\ \text{row } p \rightarrow \\ \\ \end{array} \begin{bmatrix} X_1 & & 0 \\ & \ddots\ 0\ \ddots & \\ 0 & & X_N \end{bmatrix}$$

p = element connected to receiver

$g_e(\theta)$ = element pattern factor

SLIDE 4

ARRAY OUTPUT AS FUNCTION OF TERMINATIONS

Angle of incidence = broadside
(simulation)

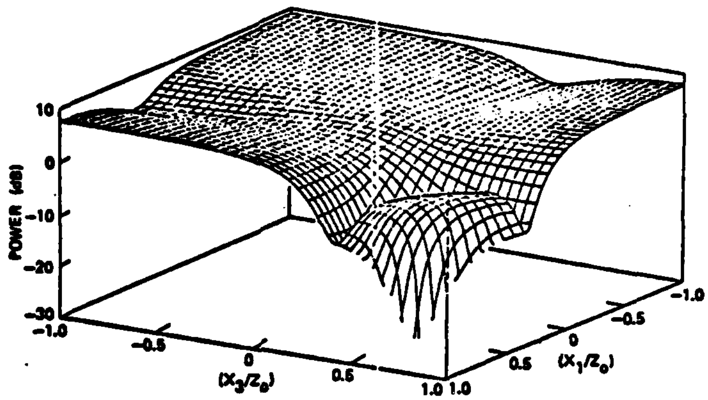
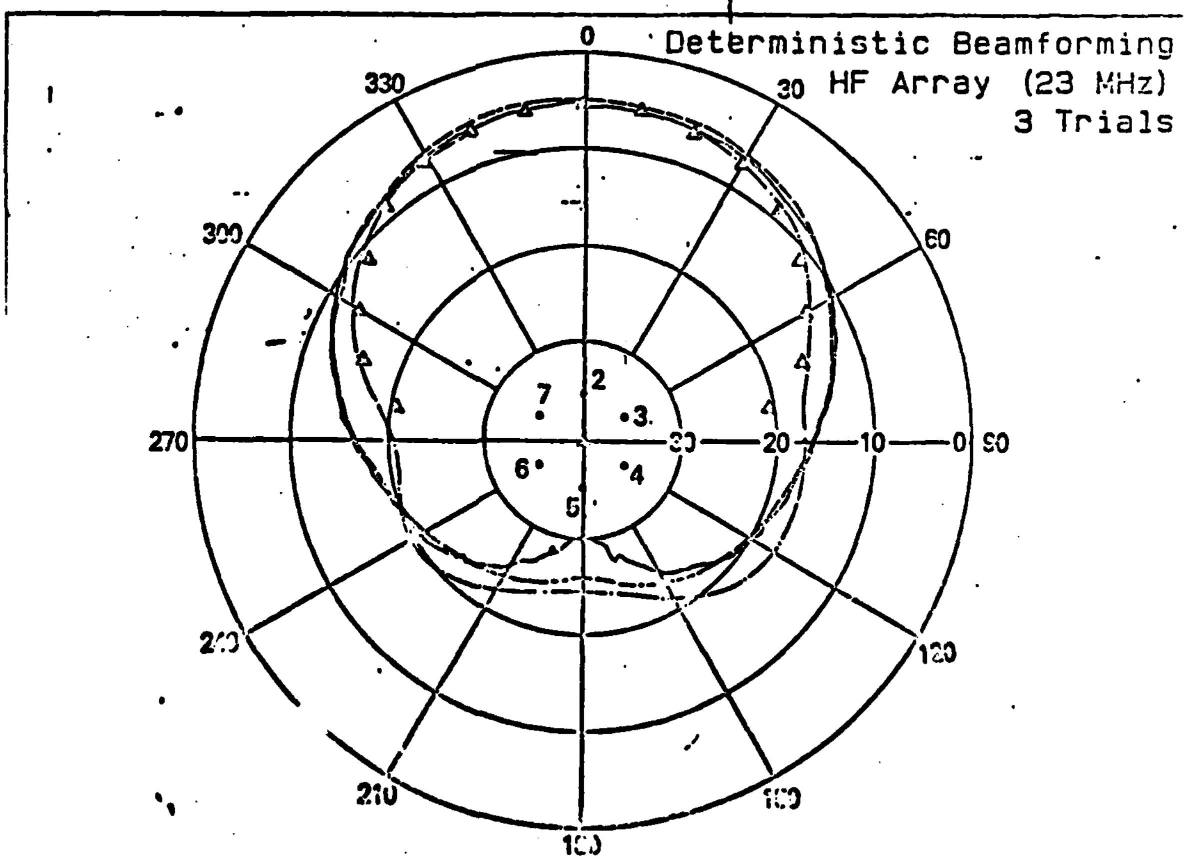


Figure 2.

SLIDE 5

5/83

Figure 3

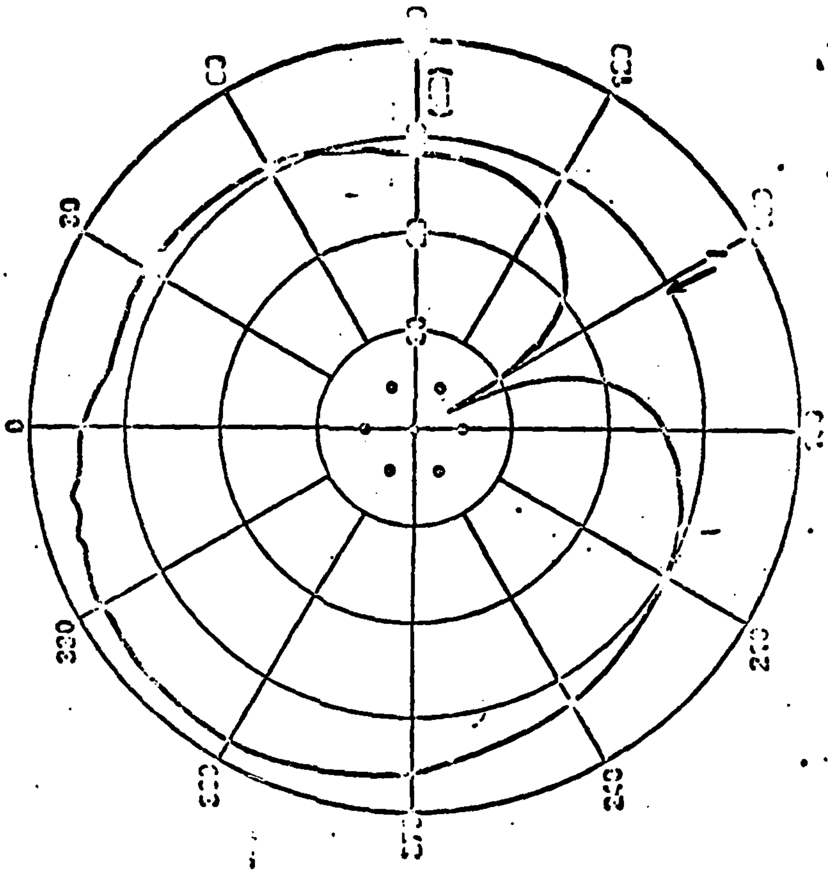


SLIDE 6

5/83

Figure 4

Manual Adaptive Nulling
HF Array (23 MHz)



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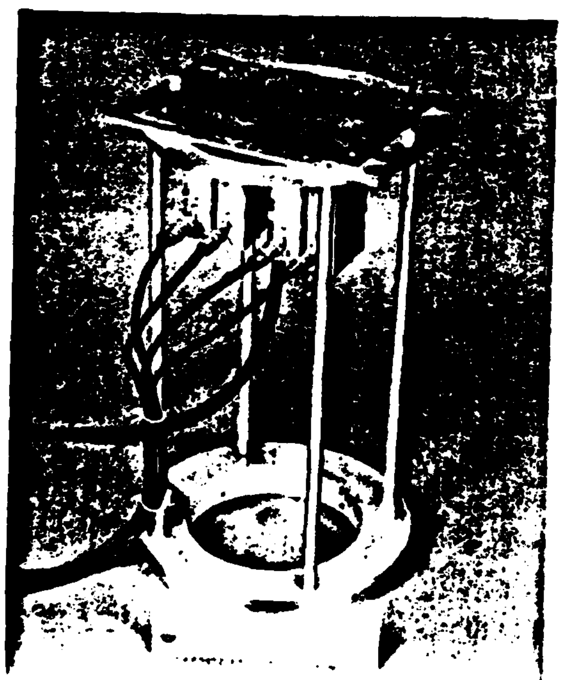


Figure 5

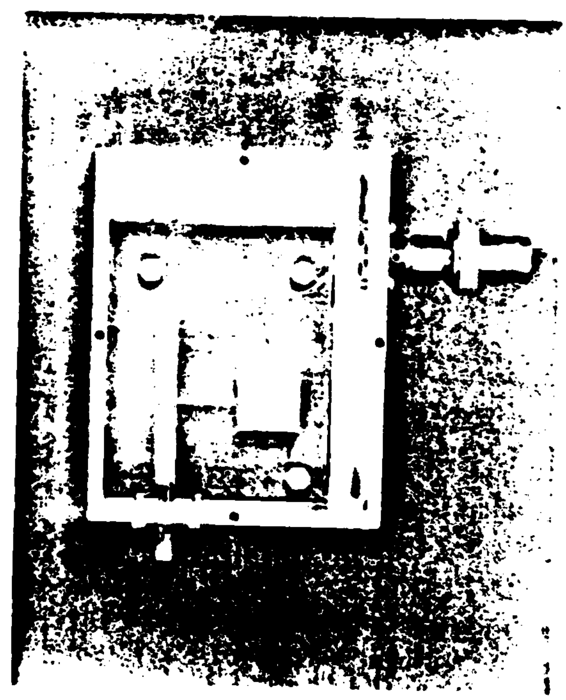


Figure 6

SLIDE 8

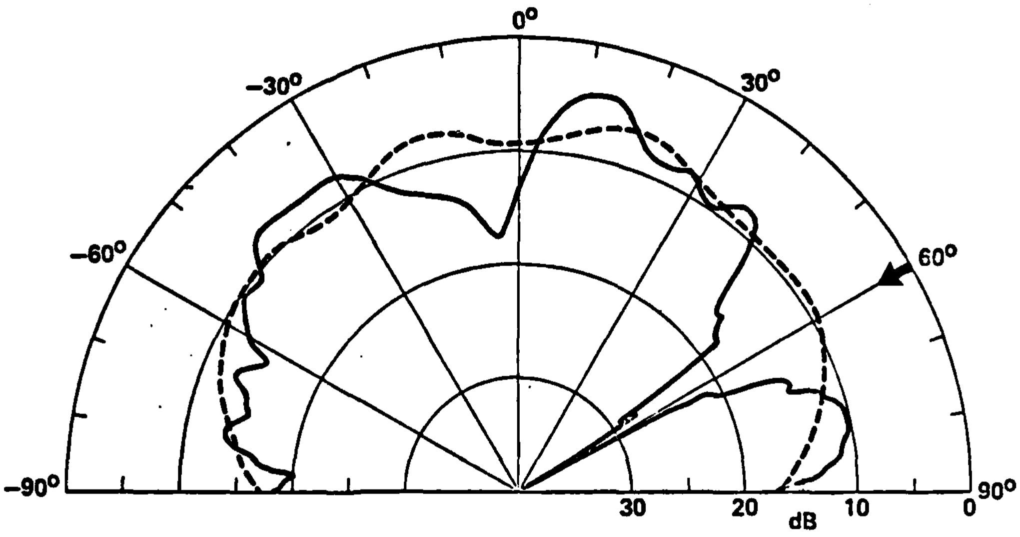


Figure 7

SLIDE 9

CONTROL ALGORITHM

Let: $X_i(j)$ = reactive load on antenna i at iteration j

Then: Steepest descent equation states that

$$X_i(j+1) = X_i(j) - K(\nabla_x \epsilon)_i$$

where K = rate constant (or feedback constant)

ϵ = criterion function

Possible criterion functions include

V_o

$\int_\tau V_o \, dt$

$\hat{S}/\hat{I}$

SLIDE 10

REACTIVE LOAD VARIATION DURING NULLING
Angle of incidence = 45 degrees
(simulation)

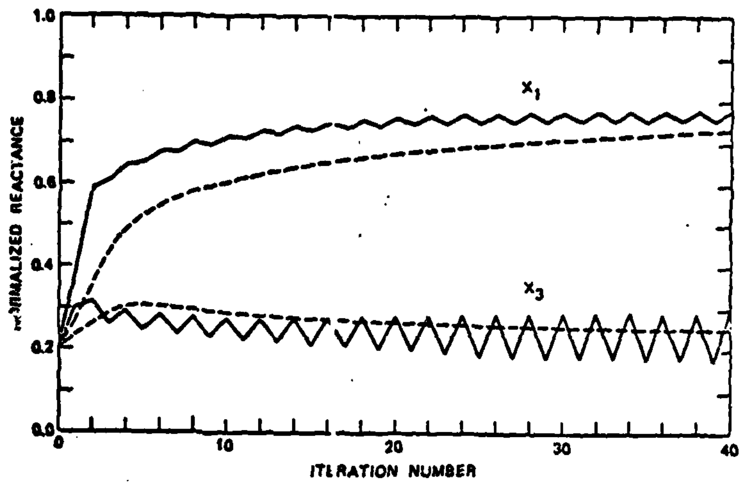


Figure 8

SLIDE 11

5/83

BEAM HISTORY
Angle of incidence = -45 degrees
(simulation)

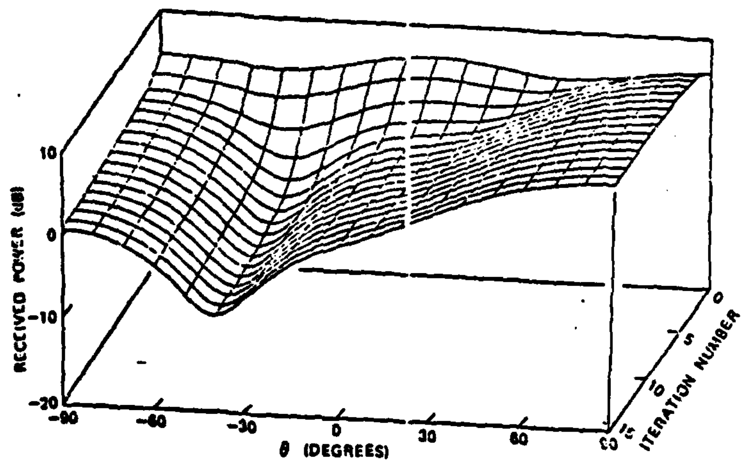


Figure 9

SLIDE 12

5/83

Array Measurement System

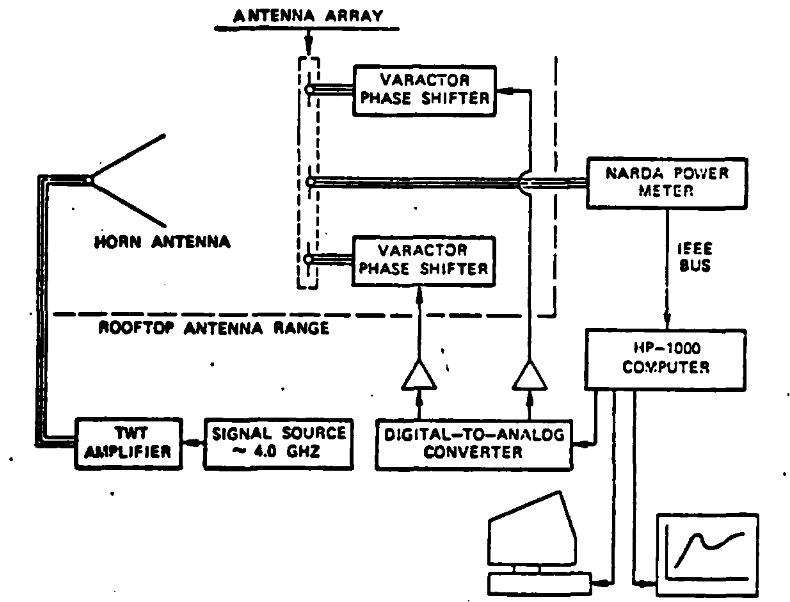
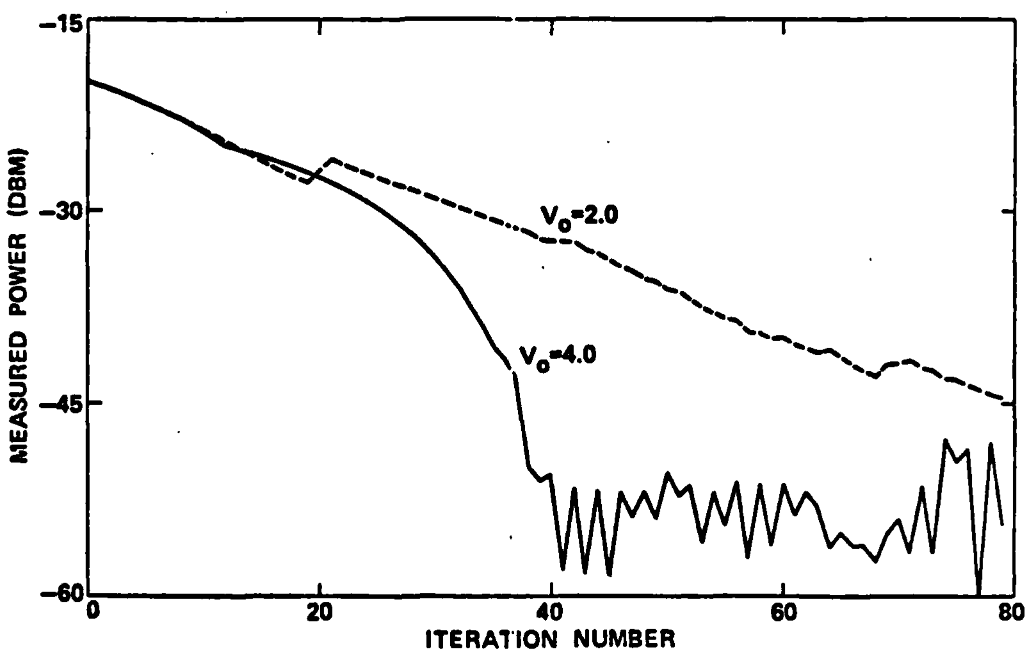


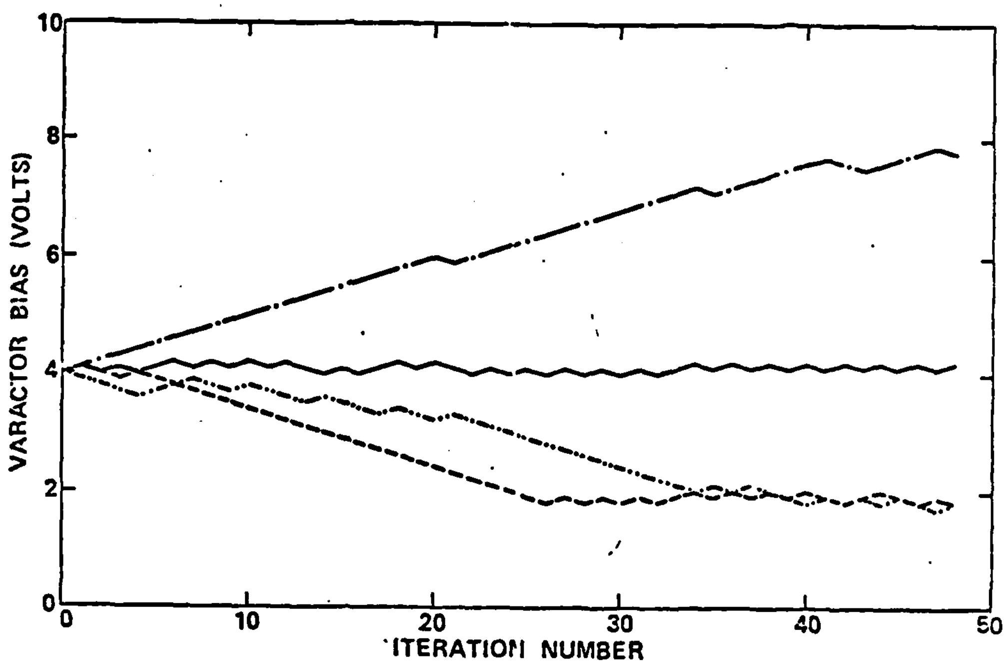
Figure 10
SLIDE 13

Interference Reduction



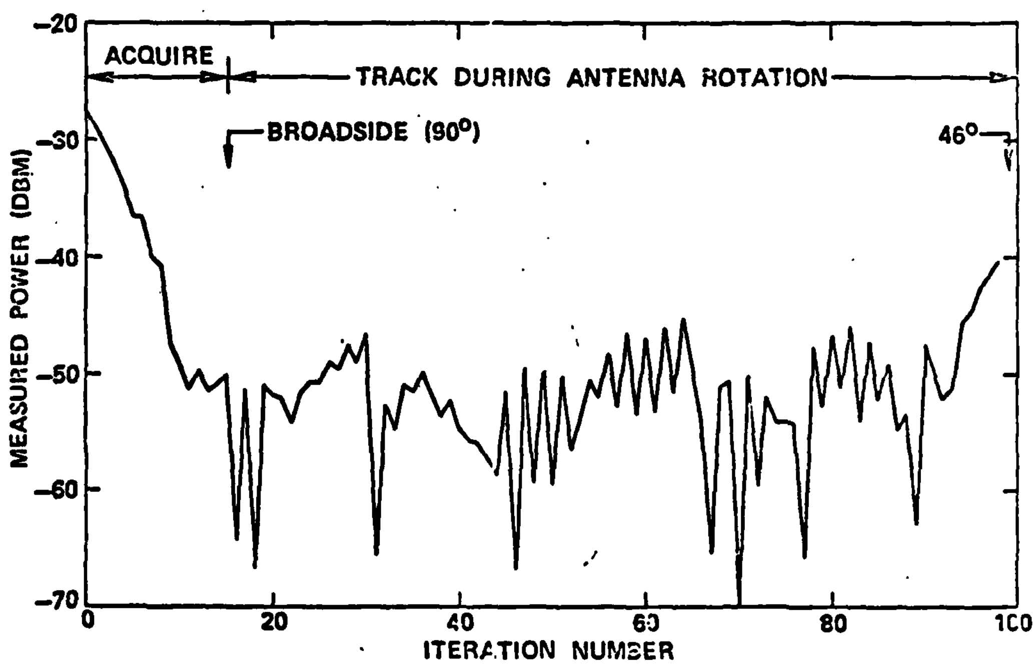
SLIDE 14 Figure 11.

Reactive Load Variation During Nulling



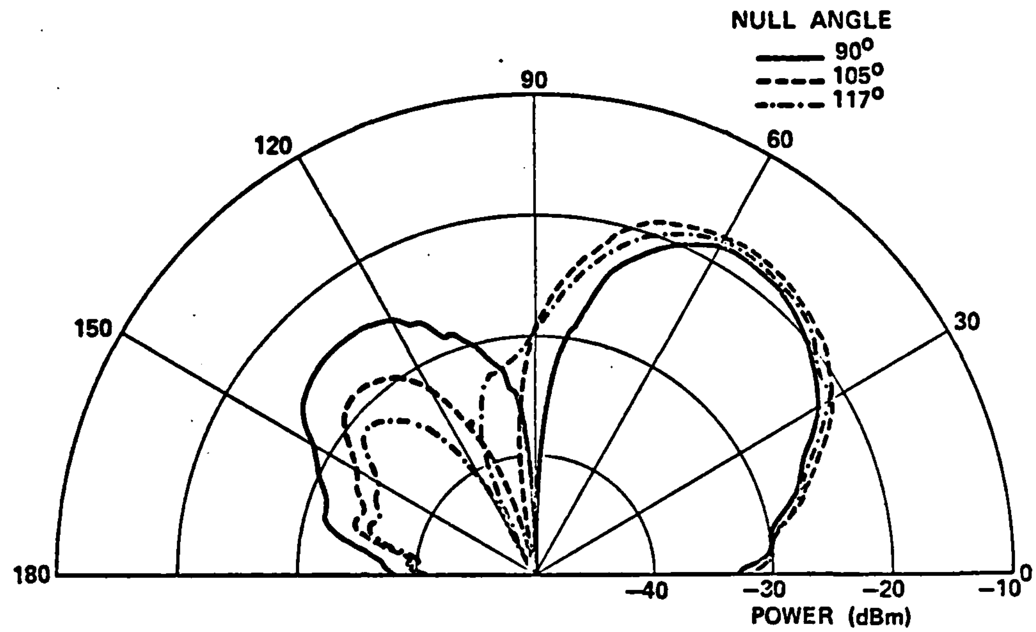
SLIDE 15 Figure 12

Interference Power During Nulling and Subsequent Antenna Rotation



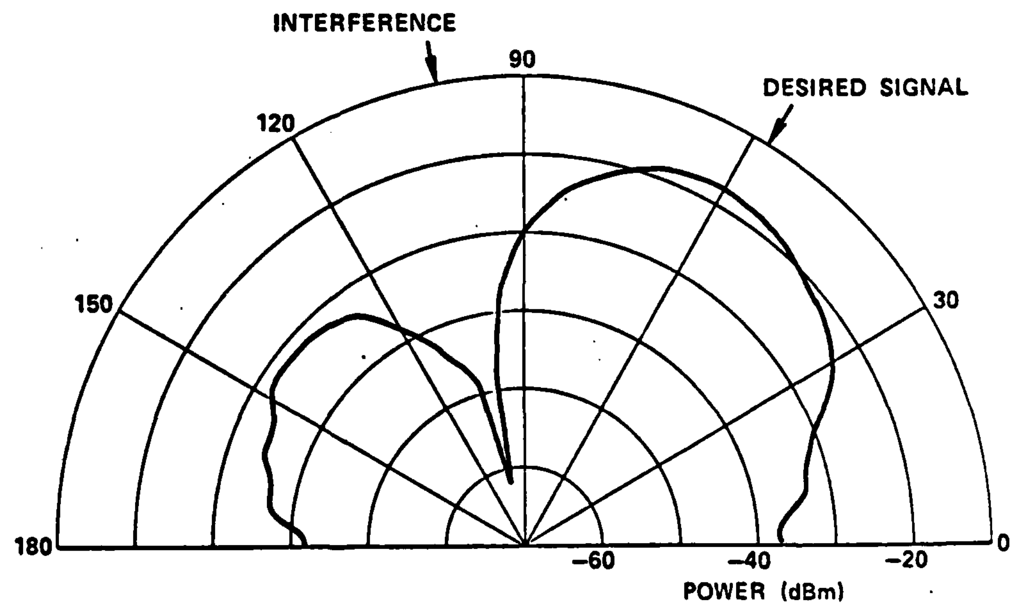
SLIDE 16 Figure 13

PATTERNS DURING ROTATION OF ANTENNA



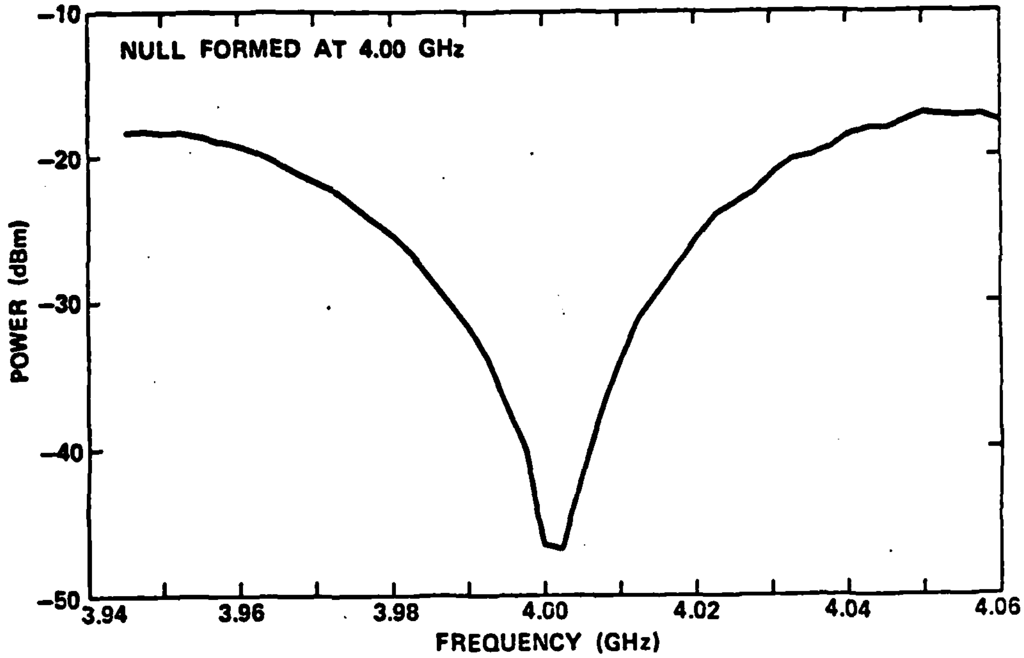
SLIDE 17 Figure 14

SIMULTANEOUS MAXIMIZATION TOWARDS DESIRED SIGNAL AND MINIMIZATION TOWARDS INTERFERENCE



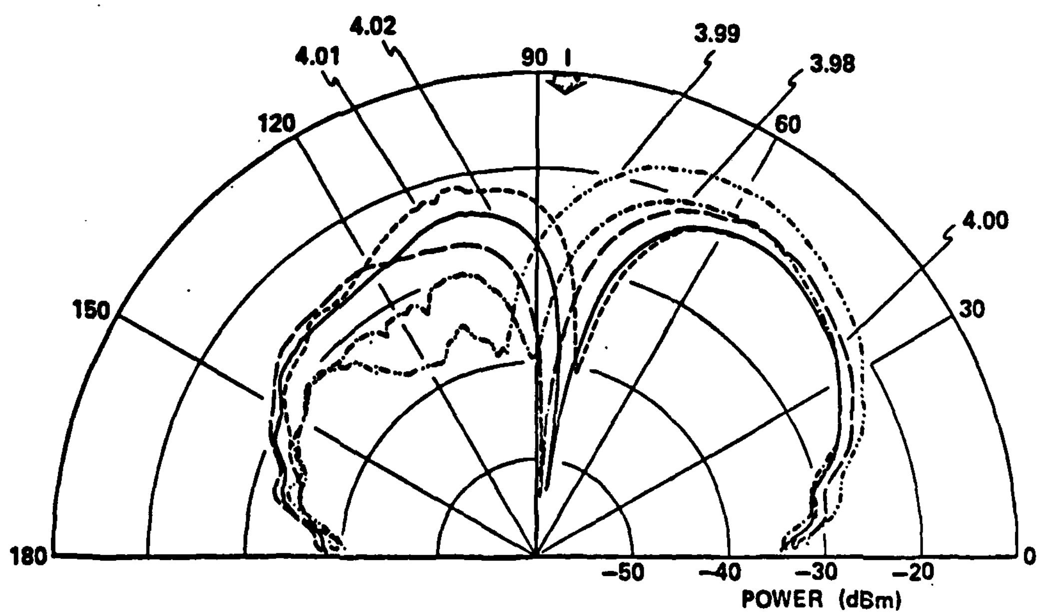
SLIDE 18 Figure 15

FREQUENCY RESPONSE



SLIDE 19 Figure 16

FREQUENCY DEPENDENCE



NULL FORMED AT 4.00 GHz

SLIDE 20 Figure 17

ARRAY SIGNAL PROCESSING IN A SPREAD ~ SPECTRUM ENVIRONMENT

DISCUSSION

QUESTION TO JOHN BAILEY: How do you keep from cancelling the signal when the jammer is of the same form as the signal. This is especially true for the frequency hopping signal and jammer.

BAILEY: I did not address any of the architecture that attempted to cope with explicit signals. My discussion was a conception of how you adapt in wideband systems, and what I showed were maximum SNR algorithms. All of those can be suitably modified using the sample matrix equivalent of the Widrow algorithm which again is under development. That is to say that if you either have an a priori pilot signal because you know something of the structure of the signal because you are already synced in a communication system. Or if alternatively you know the spatial direction of the signal in question, so that you can specify a signal direction, either of those will order suppression in the signal. To put it differently, there is no conceptual difference in the wideband system versus the narrowband system on the occurring problem of coping with the signal suppression.

LEINER: Actually there is a difference. In the case of spread spectrum, the normal assumption is that the total amount of signal power exceeds the jammer power after you've nulled, whereas in any particular sub-band that may not be the case. So when you take a subbanding approach of carving things into little sub-bands across the total bandwidth, and then you try to null in one of those sub-bands, as long as the jammer is above the signal, there is no problem, you'll null the jammer. But when you have a sub-band that doesn't have the jammer, then you are going to be nulling out the signal.

BAILEY: Well, if you assume that the jammer is jamming broadband, because you have a broadband signal, and when you sub-band, there is less jamming power coming in to a sub-band which is exactly equal to the sub-banding factors. So the J/N ratios are exactly the same in the sub-band as in the broadband.

HUTH: Is that an assumption?

BAILEY: If the jammer that was not jamming in a broadband was only jamming within a narrow band. What is your scenario?

RISTENBATT: My scenario is both signals are frequency hopping.

BAILEY: Where the jammer is also closer to you than the receiver, so that he can note the sequence of frequencies that you're hopping, so that he's jamming.

HUTH: You can't tell one from the other until you've done your sync so therefore, if your adaptation is before sync then you're out.

BAILEY: Well, that's a whole different issue than I had covered. Irving Reed will be discussing this later. I did not address the synchronization problem. It is always the synchronization problem which itself is a mechanism of avoiding signal suppression. My premise here is that you know a priori the range of the person with whom you are communicating to an uncertainty of $C\tau$, where τ is the time duration between frequency hops. That is if you have a series of frequency hops, you know the range a priori and to within that range then you know a priori the sequence of frequencies coming into the system. Of course you don't know that until you're in some sense synced up

HUTH: The problem is how to sync when you're in an environment where there is heavy jamming and in fact you are depending on the adaptive array to give you the support to get that jammer out of the way. Now you're trying to sync and you can't sync in that. So now you have a *chicken and egg problem and I don't see* how to deal with it.

BAILEY: The way to do it doesn't have anything to do with my presentation. You have to do space-time adaptation where the time part of the adaptation includes the synchronization process An example of that is the JTIDS system. How *it can be done in JTIDS will be given by* Irving Reed later. The synchronization problem is the crucial problem in an adaptive communication system. I did not address that in my presentation.

QUESTION FOR RISTENBATT: What's the impact of non-ideal signal over correlation on your approach? That is, *spurious time sidelobes of sub-sequences* and long period sequences?

RISTENBATT: (I'd like to speak to the previous question for one moment) Frequency hopping is not likely to be used without PN on the frequency hop. Frequency hopping with PN added to each hop is a very endurable spread spectrum technique for many many purposes. It is not at all difficult to do that. So I subscribe to this theory, that the frequency hop will have a PN code on it. As a matter of fact, a very decent idea is that you don't have to generate a long code. You can just have a fairly long set of known codes, let's say 100, and select the one of the 100 pseudorandomly. It is very difficult for a jammer to make progress on playing some game In effect you get the advantage of a long code, a 24-hour code, but you don't really do that. So you no longer will be fighting this problem of having this partial period autocorrelation going along. You know your code. you know it is 1 out of 100 and you lay it out. Then you have, for our scheme, basically this correlation. We still have the issue that there are sidelobes on a single PN word going into a matched filter. So we need a minimum of 2 or have to figure out some way to cope with that. We've thought of different ways. but no serious work has gone into that. So that's the way I'd like to answer that. Our favorite scheme for applying a technique like that, is that we think it applies to the FH/PN situation It will work very well in that case where you use 1 sequence out of 100 pseudorandomly. Say they'd be Gold codes, i.e. they would be nice selected codes with very good autocorrelation properties.

QUESTION: How does sub-banding reduce the effects of multipath? Aren't you throwing away multipath resolution by sub-banding?

BAILEY: It is hard to describe that *without a physical diagram*. Let me go back to this. Slide 3 Visualize 2 channels, one of which is time-delayed relative to the other. If you don't do any sub-banding you try and physically subtract the two In this example these are time-limited and so they would not be limited in the frequency domain. If I did the best I could, there would be some residue power and in fact the cancellation ratio would be on the order $(1/2)\delta/T^2$. The net result is that there is a limitation on the cancellation ratio The banddwidth in this case is 1/T. Now suppose an example four times as long, i.e., the pulse is 4 times longer and I put the same delay on. Since T is now 4T, I improve the situation by 12 dB. In effect, when you *sub-band, consider an example* of sub-banding by a factor of 4, what you're doing is taking 4 contiguous samples, and you're summing them together. For example the filter is the sum of those 4 contiguous filters. That's mathematically equivalent to a receiver

whose bandwidth is 1/4T as opposed to bandwidth 1/T by this digital summation process. So in this specific filter, the jamming residue is down 12 dB. We now talk about multipath. I'm talking about near field multipath. I'm not talking about multipath of the type where the differential delay is large compared to the reciprocal bandwidth, but rather where it is small compared to the reciprocal bandwidth. That would be the typical case on an antenna where you had near field scatters, i.e., where some of the energy was being scattered into the antenna. So I think the answer is that sub-banding helps you in the case where the differential delay is small compared to the reciprocal bandwidth which is the general case for near field scattering

QUESTION: I don't understand the deep meaning of assuming stationary jammer statistics. For example, what if jammers use FM by noise. Is that stationary? If not, what can be said about array performance?

BAILEY: One comment on that is visualize two correlated jammers, so the jammer is not a stationary process. If you have two perfectly correlated jammers, then in effect, unless you have a time delay between them, the vector sum of those is some other synthetic jammer in some other direction. So in point of fact, very often when you have non-stationary correlated statistics, things actually work better. In effect you have less adaptive degrees of freedom than you would in a pure Gaussian uncorrelated noise statistic case.

DUPREE: Were you saying in the case of the correlated jammers that you assumed that they are phase coherent?

BAILEY: Yes, or that the cross-correlation between the 2 jammers is not a zero-mean process. There is some correlation coefficient, which means they are partially coherent in some sense. If they were perfectly correlated, then the two jammers would effectively be one jammer in some other synthetic direction and one adaptive weight would take care of both of them simultaneously. If they are partially correlated, then it's equivalent to 2 jammers in some other synthetic direction, one big and one small, depending upon the correlation coefficient.

DUPREE: But very slight variations in the relative phase of the 2 jammers would cause the apparent direction of arrival to vary very rapidly, in which case you might have a difficult time adapting to the case of two correlated jammers.

BAILEY: That might be, but it would have to be very rapid indeed because in a wide bandwidth system, assuming one is using the class of honest matrix inversion techniques, keep in mind that the sample base is only 2N samples, and they are reprocessing the same data. To take an extreme case, if you had 100 MHz bandwidth, and you have 10 adaptive degrees of freedom, the time base that we are talking about is 1/5 microsecond. One would not expect phase variations over such a short period of time. In other words, the potential ultra-high speed of the adaptation process I think will save you in most cases.

SIMON: Just a comment about correlated jammers. I think if you look at the problem from an electromagnetic point of view, 2 waveforms coming from different direction can never combine across the physical aperture as if it were equivalent to a single signal, a single plane wave across the same aperture. So when one talks about two jammers, there certainly is a temporal correlation between them, but there is not a spatial correlation between. It's an over-simplification to talk about two jammers combining as if they were equivalent to a single jammer coming from a different direction.

BAILEY: I agree, you never have a correlation coefficient of unity.

COMMENT: I think what it has to do is the spatial disparity. From a temporal point of view you can have whatever correlation they have. You can have unity correlation between two signals coming *from different spatial directions.* But I'm saying that from the point of view of the processing aperture there would be the effect of two spatially disparate signals which is not the same as the effect of a signal coming from some other direction.

BAILEY: I disagree with that. We will discuss this at a later time. In fact we've simulated that. You just have the vector sums of 2 signals which are correlated.

COMMENT: But they are not equally correlated at every point across the aperture. That's my point.

BAILEY: In a narrowband system, I said, you did not have the time delay problem. In wideband systems where you do have the bandwidth aperture or a bandwidth product problem, what you say is quite true. It wouldn't be true of course, in a narrowband or sub-band.

PRICE: With respect to the question about multipath addressed to John Bailey, he avoided the situation of what is called *the distant multipath or distant scattering* where the multipath extends over more than the reciprocal bandwidth, as he put it. It strikes me that he is still in a very nice situation even there and needn't avoid it, because if I understand what he's doing with recombining all these narrowbandings coherently, I trust not incoherently, then in effect, he's essentially building a RAKE in the frequency domain.

BAILEY: That would only be true providing that the differential delay between the main path and the multipath is smaller than the reciprocal of the bandwidth of a narrowband filter, or otherwise physically the energy is not time-coincident to cancel.

PRICE: There may be problems in your adaption loops, but if everything worked perfect, I think you would simply be building a conjugate filter in the frequency domain. That's my impression, and that is a RAKE and that is fine. I think it would work. I don't see why you should feel negative about it.

BAILEY: You're building a filter based upon samples of some time base. If the multipath is somewhere else in time, then there's nothing to cancel. They have to subtend to the same time basis, in some sense, and spectrally it's equivalent to saying that the reciprocal of the bandwidth of the narrowband filter is larger than the differential time delay between the jammer and its multipath replica.

PRICE: I haven't been thinking of multipath in connection to spatial aperture. Its true maybe there are some applications there.

FEINTUCH: In spite of about 15 years of research on adaptive arrays, very few military communications systems, spread spectrum or otherwise, have incorporated adaptive arrays. Why do you believe this is the case?

REED: Most importantly, the cost and technology. I also think the field hasn't progressed really to the point where the cost of the experimental program superceeds the cost of the actual equipment. Then there are very few successful algorithms that have been developed for communications.

BAILEY: I have one comment and that is that adaptive communication systems have an inherently harder problem than active radars, namely the synchronization problem, to avoid signal suppression due to the fact that spread spectrum a priori are usually high duty

factor waveforms. That being the case, the better techniques that we are talking about are basically digital techniques, and even the digital techniques as applied to active radar systems are very recent. Everything has really been done within the last 5 years in terms of digital processing, with the exception of HF in sonar, where the interference is very low. Practically, AD converter technology, and other limitations behind broad bandwidth systems make this a very recent technology. Perhaps the other answer is just physically money. There has been a lot more money available in developing sidelobe cancellation for large scale military radar systems than for communication systems

RISTENBATT: There was one system very recently which tried to combine adaptive arrays with spread spectrum. That was the SEEKTALK. I think that the following factors are true. Retro-fitting the type of vehicles they were talking about there for the Air Force, turned out to be very very costly. I think it is also true that when they went to see the research community, when they needed the system, not all the gaps were filled. I'm not really sure that it worked as well as they hoped, but that's too dangerous to say without someone more official saying it.

DUPREE: I'd just like to comment on the question. Until recently, we did not have the payload capacity to incorporate this type of technology. We now have a 500 lb payload capacity. At the same time, we have a lot of other competing technology on signal processing which could be in for the share of the load. And unless you can produce a system as I said, which comes on the order of 100 lbs and a few 10s of watts, and which has low risk and built in a lot of redundancy, you cannot sell it to the program managers. And that's basically where it stops. If the program manager feels that this is technology he doesn't understand, it has some risks. If it is too costly in terms of weight and power, you just don't get it on the satellite.

DINGER: Let me take a crack at answering that question as applied to HF where the bandwidth of components is not so much the problem. The reason you do not see adaptive arrays in HF systems for say shipboard use is the fact that simply adding an HF array to an existing system is not suitable. You really have to redesign the entire architecture. You have to go to the new waveform, new types of transmitters, you also have intermodulation problems on shipboard. It really amounts to having to throw out the whole system and starting over new. It's kind of evolution versus revolution and you just cannot tack on adaptive arrays and really make them perform like you would hope. I think that shipboard HF has slowed down tremendously the advent of adaptive arrays in that application.

COMMENT: In the packet radio work we've done, we actually did try to build an adaptive array for the advanced wideband packet radio. The adaptive array worked just fine. The problem was when we hooked up antennas to it. I'm only being partially facetious. The real problem we ran into was coupling between the antenna elements which made all the assumptions that went into the design into the circuitry invalid. And I've seen this before in other adaptive sort of antenna technologies. So I don't know how much that has played upon in other people's work but I've seen it in two programs that I've been in.

SIMON: Irv Reed, in his historical overview mentioned Gram-Schmidt orthogonalization procedures. I've looked into that a little bit. My understanding of it is basically that it acts as a preprocessor in a fully adaptive array and the sidelobe canceller replaces the sidelobe canceller to

some extent. My question really is, to what extent is this technique applicable in the spread spectrum environment since I haven't really seen anything which has used it in that, and what are its limitations?

It seems from my understanding to have somewhat the best of worlds between achieving a better speed of convergence with respect to the closed loop systems but perhaps not quite as fast as the sample matrix inversion approach.

REED: Simulation results indicate its very close to the same speed. John Bailey has another talk on this whole subject.

BAILEY: I will make a comment. In the book by Monzingo and Miller, they talk about Gram-Schmitt only in the context of a preprocessor. Unfortunately I think that represents a misunderstanding on the parts of the authors of what is achievable with that type of orthogonalization. It is my view, and my co-workers' view that the orthogonalization technique of which the Gram-Schmidt is a type represents currently the best technique within the class of matrix inversion algorithms. It can be configured in such a completely general way that it equivalently and implicitly forms and inverts the covariance matrix, applies the steering vector, and applies the data. This is all done implicitly with a potentially feed forward network of circuits that is algorithmically exactly equivalent to doing the ensemble matrix inversion algorithm. It has all the potentials for applying VHSIC technology.

SIMON: Has it been used in the spread spectrum environment?

BAILEY: It has and it hasn't. An actual device using this was built at NRL and appears in the open literature by Bernie Lewis and Frank Kretchmer. They sub-banded in a Gram-Schmidt network, multiplexed, and adapted in 8 separate sub-bands and then recombined the data. The reason was that the receivers that they were given were so poorly matched that this offered a mechanism of matching receiver much as I outlined in the synergistic effect of receiver alignment. But it did pertain to spread spectrum in a sense that it addressed the problem of having to adapt in separate bands. The reason for adapting was that the receiver was mismatched rather than time delay and there's no conceptual difference. Many people have done simulations corroborating, if one believes in simulations, that this works. We have done it. So has Hughes, and also Lockheed, independently.

SIMON: There's another issue which sometimes comes up, and some people talk about as sympathetic nulls. I may be misusing the term but what I mean is this, in most of the algorithms that you talk about, you are maximizing the SNR, or some equivalent criterion to that. Indeed you may maximize the SNR, you null the jammer, but you also may reduce the signal by an order of magnitude. To what extent is this a problem and what is being done about it, and what kinds of algorithms solve this problem?

REED: That's part of the reason I mentioned the algorithms of Frost and Owsley. Owsley, and also Applebaum, developed algorithms which utilize constraints which preserve to a great extent the quality of the main beam so you don't lose too much SNR. A number of these have been applied and some of them work quite well.

SIMON: Does this in effect reduce your number of degrees of freedom when you do something like that?

REED: Not really.

BAILEY: I wrote a paper on this a few years back based upon what can be done experimentally on a RADC system at HF incidentally. Just to rephrase the

problem, suppose as an example you had a fully adaptive array. Suppose you preformed a beam and you wanted it to maintain good low sidelobe structures. Suppose you took all but one element in the array and performed a covariance matrix algorithm, and applied a steering vector of I on the formed beam. Well, keep in mind that one is maximizing the signal-to-interference-plus-noise ratio, and the best solution would be to untaper the antenna. In the absence of jamming you would end up with uniform distribution in Gaussian sidelobes, which would not be the solution you wanted. Applebaum wrote a key article several years ago applying main beam constraints to avoid that happening. Now what happens when you begin to add jammers, is not that you lose adaptive degrees of freedom, but when the number of jammers begins to approach let's say, 1/4 of the number of elements in the array, then the ability to maintain those constraints degenerate. Covariance matrices are linear, in the sense that if you have one jammer on and form a covariance matrix, then the steady state solutions are identical. The sum of those is identical to what you'd get in the presence of both. So this suggests the ability to take the measured steady state covariance matrix with a sampled covariance matrix, then adding synthetic covariance matrices to it as a means of constraining the system, and then adapting to that. Some people have also done that type of work. What will often happen is that the level of cancellation isn't as low as it could be, but generally that works very well because we are talking about power inversion. The relevance of that term is that if the given jammer is say 20 dB above receiver noise, then in the steady state after cancellation it will be 20 dB below receiver noise, which is overkill. If you can apply constraints you only end up cancelling 2 receiver noises as an example. That's still very good. Many of the constraining techniques have that property.

DUPREE: For fully adaptive systems you can easily plot a set of curves that show the S/N degradation due to the jammer. It turns out to be a function of the jammer's sidelobe level on a directional beam in the non-adaptive state. For example, if the jammer happened to be 10 dB down on the non-adaptive beam pointed at the user, then there would be on the order of half a dB of degradation of SNR in the adaptive state, assuming that quantization noise and so forth didn't floor you out. The maximum S/N enhancement occurs when the jammer is on the 3 dB point of the quiescent beam and in that case the maximum S/N enhancement due to going to the adaptive system turns out to be about 60 dB less than the interference-to-noise ratio. The resolution or beamwidth then goes back to the original design at the array, the resolution which you design into the array as a function of aperture size parameter.

HUTH: I have a question on rapid matrix inversion which is inspired by the Gram-Schmidt question. It called to mind something that I believe is been used in telephone equalization for digital systems, a la Lucky. The matrix is forced into a circulant and then the eigenvectors are nice sampled sinusoids. You can do the inversion very rapidly with distortion of course. So I was wondering about say using 4N samples rather than 2N and forcing the thing into a circulant. Has anyone been looking at that?

BAILEY: I've looked into that a little bit in the case of equalization networks, where the matrix is basically Toeplitz in order for it to be a circulant. The 4N x 4N matrix must be Toeplitz. Now in the case of an aperture like a linear array with equally spaced elements, for simulation purposes, one might say it's Toeplitz. But in reality, errors in the antenna in terms of

placement, limit your achievable sidelobe level deterministically, and now one is trying to cancel below that level. So unfortunately the Toeplitz assumption breaks down, which includes using a circulant matrix for spatial adaptation, but not for equalization or time-domain adaptation, where one knows a priori that the cross-correlation between tap positions is the same.

HUTH: So you think for time-domain processing, this has some merits in *speeding up processing* by adopting a circulant, for time-domain only.

BAILEY: Absolutely. Not only that, that circulant matrix can be .... its normal form, is actually a discrete Fourier transform, on each side with a diagonal matrix of eigenvalues. If that is an FFT, (and therefore it has a trivial, analytical inverse, because the inverse of the two matrices on each side, i.e. the eigenvector matrices) then the inverse is known a priori. As such one can analytically invert it. That's very significant because if you add very large number of tap delays, such as in some large PN sequence, to talk about inverting a 1000 by 1000 covariance matrix is virtually hopeless. But if you knew it was circulant, then you could do it analytically.

QUESTION: I was wondering about the statement about combining the PN with the frequency hopping. Does that imply then a slow frequency hop system?

RISTENBATT: No, that does not imply slow frequency hop system. We are talking about 3000 hops per second, so that wouldn't be slow.

QUESTION: Let me rephrase it. Are you assuming that you are coherently despreading the direct sequence, so the length of time you spend at any given frequency has to be long enough for a carrier tracking loop to lock up to allow you to coherently despread.

RISTENBATT: Why do we need a carrier tracking loop? We have an incoherent matched filter. The integration is coherent, the carrier phase is incoherent. There's a big separation. We coherently integrate over the spread spectrum code but we detect the hop incoherently.

QUESTION FOR JOHN BAILEY: Concerning the adaptation time when you sub-band. Does the adaptation time go up even though the number of computations isn't up?

BAILEY: The answer is yes, it goes up by the sub-banding factor. I showed that in one of my final block diagrams, but I was running out time at that point. So that if you had a narrow band system, comprising 2N samples, the bandwidth B sampled at 1/B intervals, then the total time base of the adaptation would be the 2N samples times the bandwidth of the system. The fact that we're sub-banding means that the bandwidths are narrower then the sub-bands, say by a factor of R. The total adaptation time is increased by the same factor. That incidentally is the reason why there's no additional computational loading in terms of computations per unit time. Because if you sub-band by a factor of 4, the time-base of the adaptation is 4 times as long, so even though you must now adapt in 4 independent bands the computations per unit time remains the same. *Whether that* bothers one depends on whether the 2N samples times the original bandwidth times some factor R, which represents the number of sub-bands is still a sufficiently small period of time in terms of such things as how often one frequency hops or whatever. The technique of reprocessing the same data however, normally makes one relatively benign to that type of thing, i.e. you cannot afford *to adapt, have a hold and then apply it to* new data. But if you're reapplying the

same information from the same data. the real question becomes, how rapidly the jamming field is changing due to frequency hopping as opposed to the time base of your adaptation.

FEINTUCH: So it's taking the same time in iterations?

BAILEY: Suppose you batch process over a period of time. Let's take explicit numbers, a 10 MHz system at 10 adaptive degrees of freedom, 2N is 20. At 10 MHz bandwidth that would be 2 microseconds. If I had to sub-band by a factor of 8 to get another 18 dB of recruitment in my achievable SNR (since it goes up by the square), that 2 microsecond time base of the adaptation will increase to 16 microseconds. Such a technique would work in the frequency hop system, providing one did not count frequencies of less than 16 microsecond intervals as an example.

FEINTUCH: There was another question about variable delay adaption rather than phase adaptation. Anyone have any thoughts on that?

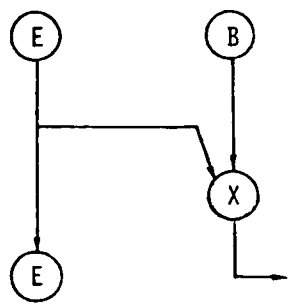
ANSWER: Suppose you have some sort of an array that is composed of elements scattered around. The real critical factor is the delay matrix between the various elements as opposed to the phase matrix because the phase matrix changes, but frequency delay doesn't. That leads one naturally to an adaptive system based on variable delay elements as opposed to variable phase elements. Somebody like to comment on that? I didn't see a lot of that in the talks.

BAILEY: <u>Slide attached</u> We know of at least two military systems that use that technique currently in hardware I can't talk about that. What we've found is that technique in principle does indeed work. It is computationally intense and requires more A-D convertors than sub-banding as a technique. The reason for that is that the total number of computations to form the inverted covariance matrix goes up as the cube of the dimension of the covariance matrix. We have two tap delays per channel, then you have a 3N by 3N covariance matrix and therefore it takes 27 times as much physical computations as we performed in the matrix, and the number of computations of unit time goes up by a factor of 9.

QUESTION: I saw the system you put up that was not the system I was thinking of. What I was thinking of was a system where basically for each element there is a variable delay element which has a lot of very good resolution, i.e. has resolution on the order of a phase-shifter. And then you'd have a magnitude if necessary for each element, and then you'd go into a sum-up and then your algorithm on that. That would seem to be the natural system that the LMS type of systems would lead you to if you were attacking a wideband signal where you have an array of ....

BAILEY: Normally what one would do with a wideband system if the direction of the signal was in a known site is time-delay steer by introducing the appropriate a priori known delays in the direction of the signal. However, the delays that you must put in to cope with jamming in the sidelobes is relative to that position. You do not know those delays a priori. Now if there's only one jammer, that is only a simple but pathological case. If you have more than one jammer simultaneously from unknown positions, then you cannot use an a priori variable delay, and in fact you may have to have two deterministic taps, and then make the whole system adaptive using those tap delays.

(3) SINGLE CHANNEL SIDELOBE CANCELLOR



$$B_A = B + wE$$

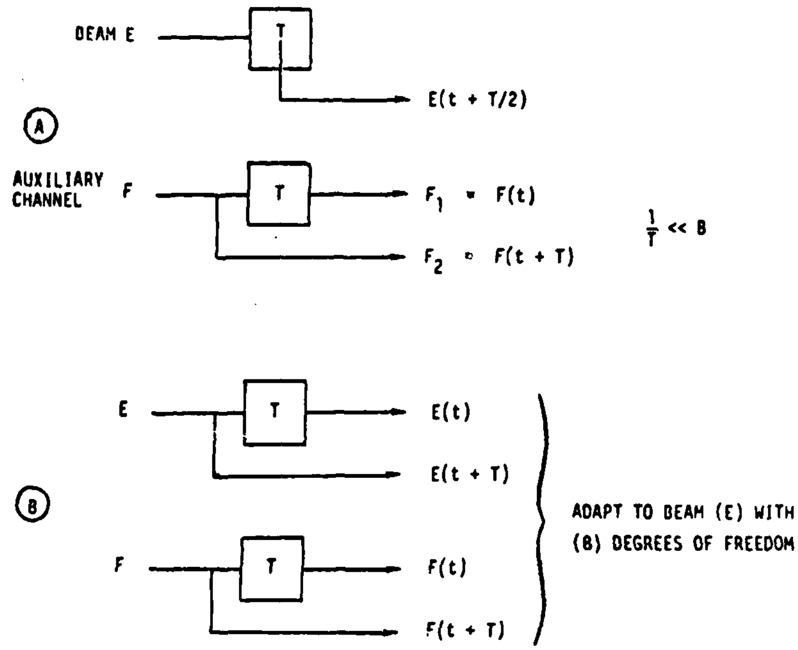
GRAM SCHMIDT ORTHOGONALIZATION

$$w = -\frac{\widehat{B E^*}}{|\hat{E}|^2}$$

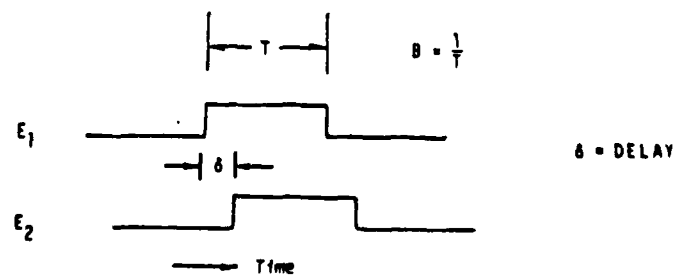
$$\widehat{B_A E^*} = 0$$

SLIDE 3

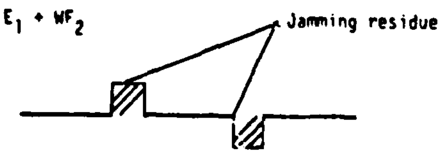
TAP DELAY IMPLEMENTATION



TIME DOMAIN SLC LIMITATION



Eg 1 $(E_1) = E_1 + WF_2$



SESSION 2 - SPREAD SPECTRUM COMMUNICATION IN JAMMING

GAYLORD HUTH

This session is Spread Spectrum Communications in Jamming. We are going to have Bob McEliece talk first, followed by Barry Levitt, Pravin Jain, Jerry Gobien, and Seymour Stein. Bob McEliece and Barry Levitt are going to tell us about the game min-max design, or how the communicator beats the jammer (or vice versa) from an analytical point of view. Pravin will tell us about the system aspects of satellite communications using spread spectrum to combat jamming. Jerry will talk about the tactical world, and system design with jamming. Seymour is going to try to tell us what all happens after all that occurs.

Bob McEliece graduated from Caltech long ago, and worked for JPL for many years. He went to the University of Illinois at Urbana for 4 years, and now he has *come home to the Caltech EE Department*. so Bob, you want to try this?

BOB McELIECE

Today I want to work through a specific example of one "jamming game" which may have some applications, and prove to be interesting. My general view is that after the processing gain is given, and after the antenna has partially nulled out the jammer, he's still there to some extent. However, you still want to get a signal through.

From this abstract viewpoint, I have a simple block diagram of the jamming game. See Figure M-1. There are 2 players, the transmitter and the jammer. I denote the transmitted signal by X and the jamming noise by Z. I've also assumed that X and Z are independent, so I'm not talking about follow-on or repeat jammers. It would be interesting to talk about jamming signals which are somehow correlated with the transmitted signal, and where you can measure correlation or mutual information or something, but I don't know any very interesting results about that problem. These two signals, one is noise and one is signal, are somehow combined and the receiver gets a signal which I denote by Y. Of course, the transmitter wants the received signal to look like what he transmitted, while the jammer has the opposite intention.

Anyway, this is a *game* where I model the signal and noise stochastic processes or maybe just a random variable. We have two players, and we need to have some rules, or at least we need a referee. Because we have a game, we also need a payoff function, to tell who won the game, or the score of the game, anyway. Normally the score would be received SNR or bit-error probability. But since I'm an information theorist at heart, and also because there are some nice mathematical properties of this function, I use the mutual information between the random variable X and the random variable Y as a payoff function. Mutual information measures the channel capacity, the best you can do with arbitrarily complex coding. So generally you can only justify this particular payoff function when you're also talking about a coded antijam system, because without coding, you can't exploit the full channel capacity. You really need coding, if only in the form of diversity, to get what's promised or some of what's promised.

Basically, I take a Von Neumann approach see Figure M-2 and hope that there is a saddlepoint (there usually is) when mutual information is the payoff. The signaller reasons as follows. If I use strategy X, the worst that could happen is that the jammer will discover my strategy and choose his countermeasure, his

jamming strategy Z, so that my channel capacity will be as small as possible. That's a conservative view point. I use X, and I assume that there will be some crypto-variables or something that he won't be aware of. But he'll know everything else. He knows my modulation parameters, he knows my power, he knows everything. The worst among the possible strategies that I presume are open to him is to choose the Z that minimizes this payoff function. I will in turn choose the X for which the minimum is a maximum. That's the max-min. The jammer takes the opposite, dual approach. That is, if I (the jammer) decide to use jamming strategy Z, the worst that could happen is that the communicator would discover it and in response choose a strategy X which makes the channel capacity as large as possible. So I will choose that jamming strategy Z for which the maximum is a minimum. That's the min-max.

So the max-min is the signaller's value of channel capacity, and the min-max is the jammer's value. Both players take very conservative views of the world, that everything he uses will be discovered. The jammer assumes that his opponent will be Shannon himself, and the signaller assumes that his opponent will be Kolmogorov himself. That's nice analogy on several levels. In some cases (Barry will give an example) these two values are different and you get sort of an instability. But mutual information is a saddle-shaped function. It's convex one way in the signaling strategy, and convex the other way in the jamming strategy. This implies in a general setting, (if you have compactness and various other things which you would normally apply to real situations) that if mutual information is the payoff function, the two values are the same.This is a miraculous consequence of Von Neumann's theory. It's amazing that these two conservative strategies both lead to the same answer. Moreover, not only do the two players agree on the value of the game is, they have saddle-point strategies. They will both decide ahead of time, (Shannon and Kolmogorov) even though they perhaps are not willing to announce it, to use saddlepoint strategies. They assume that their opponent is equally smart analytically, and capable of making this computation. By the game theory approach, they don't care if the opponent discovers this optimal strategy. If the jammer uses saddlepoint strategy Z_0, then the best channel capacity that the transmitter could possibly give to himself is this number. Furthermore, if the signaller uses any other strategy but the optimal one, he'll do worse, i.e. smaller channel capacity. So if the signaller uses a particular strategy X_0, and if his opponent the jammer uses the saddle point strategy Z_0, then this is the channel capacity that will result. If the jammer does anything else, he will do worse.

Let me give an example, using an abstract mathematical view of non-coherent MFSK. Figure M-3. So here we have a mathematical abstraction of a frequency hop system, in which M=8, so there are 8 signaling tones. Now these may be jumping around in frequency, but let's travel with the hopper so that we can see what's going on. In every unit of time the signaler transmits energy in one of the M-possible tones. In this case one signal would be worth 3 bits of information. So that's a possible signalling strategy. I haven't said what the random variables involved are. I want to think of the most general possible jamming strategy, which depends upon the receiver structure. I'm assuming we just have a non-coherent energy detector where the receiver looks at each of the tones and detects the amount of energy, and there'll be some if the jammer is there (there's no background noise in this analysis). There will be M different positive numbers that come out of the receiver, one for each

frequency. I model these numbers as random variables. Some are bigger, some are smaller. In Z_3 Figure M-3 there is no jammer there at all. That's the most general jammer type for that particular receiver structure, the most general type of jamming strategy that I can imagine Of course not every such jamming strategy may not be implementable

Now, a broadband noise jammer fits this model, in which case the Z_i's would be i.i.d Gaussian random variables. A partial band noise jammer fits this model, in which the Z_i's would be i.i.d Gaussian with probability p which is the duty factor, and zero with probability 1-p. A one-out-of-M tone jammer fits this model, in which case the r.v.'s will all be zero, except for one which will have a given value. In fact, given this receiver, any kind of jamming strategy which you can imagine, will fit this model.

Now I would like to bring in a little twist. At the moment, the signal strength is the constant λ, which in this case is symbol SNR. But there is also some kind of an average power constraint on the jammer. The mathematical formulation of this is a bound on the mean square of the signal that he's allowed to transmit. Of course, he may have all his energy located somewhere else entirely, but that would be a waste. He's supposed to know what you're looking at, and presumably he'll put all his power in there. So in fact, the amount of energy available to the signaller is less than or equal to λ, on this scale of normalization. See Figure M-3 If we have a channel like this, and the signal is equally-likely to be in any one of these M places, we get an M-ary symmetric channel. We get a certain probability of not getting through correctly. All of the other signals are equally likely under some mild and reasonable assumptions. In that case the R_0 people and the capacity people agree that the figure of merit for this channel should be the error probability. So I won't raise, at the moment, the issue of whether capacity or R_0 is a better figure of merit. On a symmetric channel, everyone agrees So for a given parameter λ, what is the distribution of Z's, where Z is restricted to satisfy $E(z^2)=1$, but which maximizes from the jammer's view point the error probability? Although I don't think this problem has been answered in this generality before, you won't be surprised to know the answer. That is, if the signal strength is a constant, then the optimal strategy is a one out of M-tone jammer Figure M-5 If X is a constant, then the optimum distribution of Z's has at most one of its components non-zero The value of that component should just be a teensy bit bigger than the signal, so that, except for one chance out of M, your jamming tone will have larger energy than the transmitted tone and the receiver will make an error. Of course, this result goes back to Sam Houston, Barry Levitt has looked at this too, and we've made a small contribution to this ourselves since we've considered the most general possible jammer. But one is surprised to find that this is the result. see Figure M-4 The error probability for sufficiently small SNRs is a constant (M-1)/M, and then becomes the famous inverse linear function where the SNR gets a little bigger

I don't like this curve Figure M-4 because I don't like non-continuous derivatives. Looking at this curve, something always struck me funny about a system in which there's is a whole range of positive SNRs for which we have zero channel capacity. I thought about this for a while, and there are various solutions. I think of it as a genuine paradox having to do with an instability in the model. One way around this is to go back to this model Figure M-1 and say, "Now wait a minute, I started by talking about the jamming game, and min-max and max-min. But I didn't give any min-max or

max-min. The signaller didn't get to play this game. The signaller was just transmitting X all the time." Well, he was sort of playing the game if you allowed the distribution of these M frequencies to be random variable. So there is this possible game. You can imagine a non-uniform distribution on the tones, but that's silly because he wants to be uniformly distributed. It occurred to me one day that if the jammer is allowed to do pulse and partial band strategies, then why don't we don't let the signaler play this game, too? Suppose the amplitude or the energy transmitted at a particular tone was not a constant but was a random variable, subject to an average energy constraint. I think that's an interesting possibility. Then we really do have a game, because we have a random variable (X) here and a random vector (Z) here. The random variable X is subject to a mean square constraint, and the random vector Z is subject to a mean square constraint. Both players agree that the channel error probability is the payoff function. The question is, "What should they do?" This is the final result that I'm going to talk about.

So we let X be a random variable, with expectation $X^2=\lambda$. There are saddle-point strategies, and they look like this Figure M-5 They are a little surprising, we haven't seen this before but I've been telling people about this for a while. If the SNR is big, then the optimum distribution of signal amplitudes not only should not be a constant, it shouldn't be any finite number of levels. The saddle-point strategies for the signal level should be a uniform random variable, on zero to twice the energy that you are allowed. A uniform random variable X^2 on zero to 2λ will have the right expectations λ. Now what is the jammer? The jammer should be exactly the same. The jammer should also be uniform on zero to two λ. That's what he wants to be, but he can't because he doesn't have that much energy. If the SNR is bigger than 1 the mathematics says that the jammer should be uniform on zero to 2λ, with a certain probability, and zero the rest of the time In other words, a pulsed jammer, and we're familiar with pulse jammers being the optimum. What he does when he's on the air is to be a uniform random variable. You can prove this once you've guessed it or come up with it somehow: it's not hard to prove There's a similar result for low SNRs except now that for low SNRs where the jammer has more energy than the signaller, the jammer should be uniform. The jammer never wants to put energy in more than one of the M-tones, but in the tone that he chooses his energy should be uniformly-distributed. The square of the signaller's energy should be uniform and now the signaller should pulse He should be uniform at the same interval except that he has to go off the air once in a while

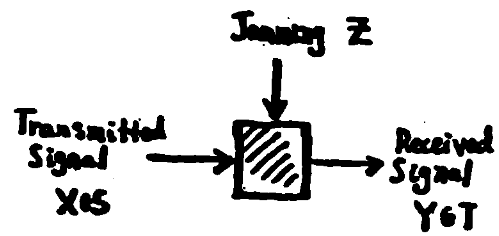
Let me just show you what it works out to be Figure M-4. This is why I said that it was some kind of an unstable model, (Sam Houston's optimum tone jammer). If you allow the signaller this extra degree of freedom, which I claim is reasonable under certain circumstances, the instability goes away The curve looks like that It's a continuous curve, it's still inverse linear beyond the same SNR and it degrades gracefully - it has a continuous derivative But perhaps more interesting from a systems, or an applications viewpoint, is that you also get a gain by letting the signaller do this, which is exactly 3 dB in this instance This is uncoded, no diversity, this is just for free, provided you randomize the amplitude The receiver doesn't have to know what the amplitude is, he's just detecting the presence or absence or a pulse or energy at each one of the M-tones. You get 3 dB for free out of this and I think that's pretty interesting. It raises a lot more questions

than it answers. So I think it's a good place for me to stop.

REFERENCE

McEliece, R.J. and E.R. Rodenich, "An Abstract View of Optimal Jamming vs. Noncoherent MFSK", paper to be presented at MILCOM '83.

The Jamming Game



Payoff Function $\phi(X,Z) = I(X;Y)$, mutual information
"channel capacity"

FIGURE M-1

Von Neumann Approach

Signaler: $\max_X \min_Z \phi(X,Z) = C'$

Jammer: $\min_Z \max_X \phi(X,Z) = C''$

$$C' < C''$$

But: ϕ is convex $\cap$ in X, convex $\cup$ in Z $\Rightarrow C' = C''$, the value of the game.

$$\phi(X,Z_0) \leq \phi(X_0,Z_0) \leq \phi(X_0,Z)$$

"Saddlepoint Strategies"

FIGURE M-2

The graph, for $M=2$.

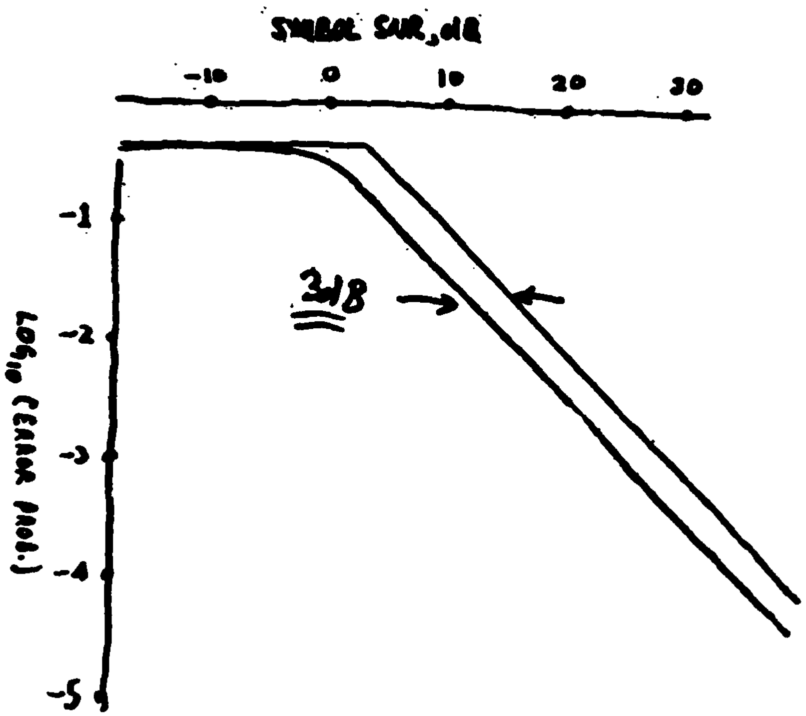
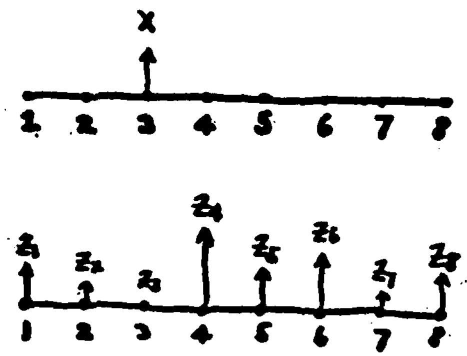


FIGURE M-4

Example (MFSK/noncoherent detection)

Signal:

Jamming:



Restrictions:

$$\left.\begin{aligned} E(X^2) &= \lambda \\ \frac{1}{M}\sum_{k=1}^{M} E(z_k^2) &= 1 \end{aligned}\right\} \lambda = \text{symbol SNR}$$

FIGURE M-3

A Result:

If X is constant $(\sqrt{\lambda})$

$\underline{Z}$ should have at most one nonzero component $(\sqrt{\lambda+})$

$$P_E = \frac{M-1}{M}, \quad \lambda \le M$$

$$= \frac{M-1}{\lambda}, \quad \lambda \ge M$$

(Hasbon, Levitt, us too.)

FIGURE M-5

Result

If X^2 is a random variable $E(X^2)=\lambda$, Saddlepoint strategies exist:

$\lambda \ge M$: X^2 should b uniform on $[0, 2\lambda]$

$\underline{Z}$'s non zero component should be

uniform on $[0, 2\lambda]$ $\quad p = \frac{M}{\lambda}$

zero $\quad p = 1 - \frac{M}{\lambda}$

$\lambda \le M$: $\underline{Z}^2$'s nonzero component should be

uniform on $[0, 2M]$

X should be uniform on $[0, 2M]$ $\quad p = \frac{\lambda}{M}$

zero $\quad p = 1 - \lambda/M$

FIGURE M-6

GAYLORD HUTH

The next speaker is Barry Levitt Barry received his bachelor's from McGill University in 1965, his Ph.D. in 1971 from MIT, and then he went to JPL, and has not been able to get out of there since. He started out in optical communications, and now he's found the same game that Bob has found, so he's working in anti-jam military communication networks now, and he's going to tell us some more about this jamming game.

BARRY LEVITT

As Bob has already mentioned to you, there are some cases of interest that are non-pathological for which there isn't a saddle-point: that is, the min-max and max-min solutions are not the same One of those cases is the familiar scenario of frequency-hopped, M-ary frequency shift-keying with diversity in partial band noise. A reference for this is the Viterbi-Jacobs paper of 1975. Just as an example, see Figure L-1 I drew a very simple-minded diagram of a diversity-5 system, with 2 consecutive symbols from a higher order alphabet (a 2 followed by a 1) being transmitted. The jammer chooses to concentrate its power only in a small portion of the entire spread spectrum bandwidth, with frequency duty factor, ρ. In this example, the diversity is achieved with fast frequency hopping, and an appropriate amount of interleaving.

Some assumptions are associated with the work by Viterbi and Jacobs Figure L-2. There's no thermal noise outside of the jammed region, and we can make perfect decisions on the transmitted signals. Each chip is independently jammed with probability ρ: if it falls in that jammed band, then it's hit and has power spectral density N_0/ρ. If it falls outside of that region, there's no noise at all. We are also going to assume perfect jamming state information, and we're not going to worry about how we derive that right now. Consequently, as a result of these first three considerations, an error can only occur if all m of the diversity chips that are used to transmit a particular M-ary symbol are jammed: that is, they all lie within the jammed region. The detection metric is going to involve noncoherent energy detection of each chip In the case where all chips for a particular M-ary symbol are jammed, the decision will be based on the linear combination of all of the energy detector outputs for each of those chips. In the absence of diversity, (diversity 1) we can compute exact error rates. In the case where we have diversity, it is mathematically more convenient to use the union bound to reduce the problem from an M-ary signaling set to a binary signaling set, and then to simplify the diversity problem by using the Chernoff bounding technique. There are some inaccuracies involved in that approach, yet it does produce closed-form solutions, which provide useful insights into the interaction of the various system parameters.

Now for those of you who are familiar with the paper by Viterbi and Jacobs, the union/Chernoff bound has this particular form Figure L-3. The important idea is that it is a function of the diversity m, the duty factor ρ, the SNR (E_b/N_0), and the alphabet size M. From the communicator's point of view, what we really want to do is to look at the minimization over the diversity m of the maximization over ρ of the bit error rate (BER) bound. The result that they found was that they indeed produced an exponential relationship between the bit error rate and E_b/N_0. The optimum diversity is the log to the base 2 of m times E_b/N_0 divided by 4. Some graphical examples of the min/max solution will be shown for the special case of m=16, so that the optimum diversity is just E_b/N_0. The worst case duty factor ρ turned out to

be 3/4; people were bothered by this because it is independent of E_b/N_0 and m. The feeling was that perhaps in the exact case (instead of using the union/Chernoff bound) ρ would actually vary a little about that value of 3/4.

Now let's take a look at the min-max problem. I used my IBM P.C. to plot the bit-error rate upper bound as a function of ρ with parameter diversity m. Figure L-4. In this case I set E_b/N_0 to 10 dB, so the optimum diversity was 10. If you take any value of diversity other than 10, then the jammer can choose ρ to give a bit-error rate that is higher than the min-max solution of Viterbi and Jacobs. So if you use m=10, you guarantee that no matter what the jammer does, you can never do any worse than this particular bit error rate. Now the implications of the curves I mentioned no matter which diversity m the communicator chooses, somehow the jammer is privy to that and subsequently chooses the value of ρ that gives the highest bit-error rate. This implies a jammer advantage. However, it is a worst case approach if ρ cannot be monitored, and that's a key point. We're assuming here that the communicator cannot measure ρ and the jammer cannot determine m As I mentioned choosing the value $m=m_{opt}$ insures that the communicator has a guaranteed maximum bit-error rate. Figure L-5

At one point, I wasn't aware of the Von Neumann criteria, and instead of computing the min-max solution, I somehow computed in max-min. Figure L-5a What I was expecting was a saddle-point solution. Suppose I plot bit-error rate against m this time with ρ as the parameter. From the jammer's point of view, what I would have hoped for is that the ρ=3/4 curve would have a minimum at m=m (opt) (as given by the Viterbi-Jacobs solution), and any other value of ρ would produce a minimum bit-error rate which is lower than that particular value. That's the result you'd find if you had a saddle-point. However, again using the IBM P.C., this is what the curves actually look like Figure L-6. You can see that if the jammer chooses ρ=3/4, you can exploit that: for example, you can go down to a diversity of 1, in which case you'll get better performance, or, for any value of ρ other than 1, the bit-error rate will ultimately go to *zero* for large enough values of diversity.

Now of course this raises some complexity issues. In a practical situation you may not be able to use a lot of diversity, but to the extent that you can, mathematically at least, Figure L-6 says that the jammer should choose ρ=1. That's the only curve for which the bit error rate monotonically increases with diversity. So what we have from the jammer's point of view (the max-min approach) is that he should use ρ=1, and the best that the communicator can do in that case is to use diversity 1. Figure L-6a From the jammer's point of view, this is the largest guaranteed minimum bit-error rate

Let's see how the min-max and max-min results compare Figure L-7. Because the max-min curve is the diversity 1 case, we are comparing an exact result with an upper bound, and the upper bound is pessimistic by about 1 dB, perhaps 1.5 dB, in this particular case. So consequently, even though there's a 3 dB separation between the curves shown, there might only be a 2 dB separation in actual performance. That doesn't *trivialize* this result because what it's saying is that if the jammer chooses ρ=1 and the communicator chooses $m_{opt}=E_b/N_0$ for the case of 16-ary signaling, the two results are going to be fairly close together Of course, for other values of ρ and/or diversity the performance can fall outside of the region bounded by these two curves

Let's now take a look at some of the risks implicit in trying to apply game-theoretic techniques to this problem. As a reference point, we have the min-max curve with diversity $m=E_b/N_0$ for 16-ary signaling and $\rho=3/4$. Figure L-8. Now suppose the communicator reasons as follows: "I know the jammer is going to use $\rho=3/4$. Why should I use diversity E_b/N_0? Let me use diversity 1. It's a simpler system; it's cheaper." So he goes to an $m=1$, $\rho=3/4$ curve. We've already seen that it will improve his performance somewhat (by approximately 1 dB). What penalty does he pay for this modest improvement? The risk is that if the jammer suspects, or is somehow able to monitor, what the communicator has done, namely that he's using no diversity or coding, he can change to a very small value of ρ. In particular at a bit-error rate of 10^{-5}, if he uses a ρ of the order of 5×10^{-5}, the communicator can end up losing about 32 dB. It's a very bad trade to gain 1 dB at the risk of *losing 32 dB if you're* discovered.

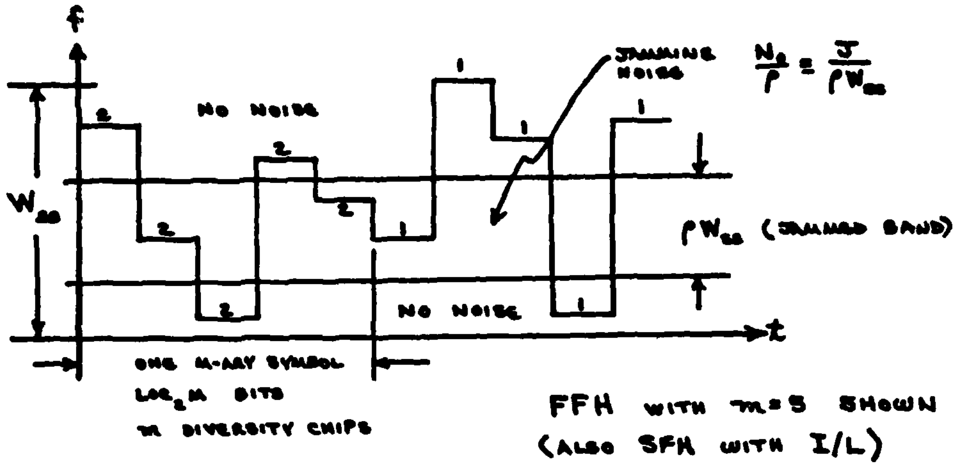
What if we go in the other direction? Instead of going to diversity 1, what if we opt for a large diversity? Again, the standard min-max curve is shown as a reference Figure L-9. Instead of using $m=E_b/N_0$, let's use $m=100E_b/N_0$. Now the diversity is not going up by a factor of 100 because E_b/N_0 is a lot smaller for the same bit-error rate. So, again at our benchmark 10^{-5} bit-error rate, in the case of the min-max solution, $m=13$ is sufficient to achieve that particular bit-error rate. On the other hand, if we use diversity 40, which is only a threefold increase, we can pick up about 15 dB. In fact, we could have used even more diversity, which would have improved the performance further. However, if the jammer ever spots that and reverts back up to the case where $\rho=1$, he's going to really zap you; that's in fact what happens here. In this particular case, if the *jammer* discovers that you've gone to a diversity which is 100 E_b/N_0 he can go back to additive white Gaussian noise, (full band) jamming, resulting in a 15 dB degradation. Of course, again, for larger values of diversity, these two curves simply diverge a little bit more, that is, the potential improvement and risk both increase. So there is a risk in trying to play that sort of game.

What about the jammer's perspective? What should the jammer do? Here's the standard min-max solution again with $\rho=3/4$ Figure L-10. We've already seen that with $\rho=3/4$, if the communicator goes to diversity 1, he'll pick up a little bit of performance. And if he goes to a diversity that is very large, he'll pick up a lot more. So with $\rho=3/4$, the jammer effectiveness can be undermined. What happens if the jammer uses the value $\rho=1$ that I'm recommending? If he uses the value $\rho=1$, he has an unexciting broadband noise jammer, but he's playing it safe. If he uses $\rho=1$ rather than $\rho=3/4$, and the communicator persists in using $m=E_b/N_0$, the two performance curves are fairly close together. So the jammer effectiveness is degraded only slightly. On the other hand, if $\rho=1$, the best that the communicator can do is to use no diversity, and all the jammer will lose is 1-2 dB, but he certainly is not going to be subject to a large amount of exploitation. So again, what we've got is a guaranteed minimum jamming effectiveness and communication performance if the jammer uses $\rho=1$ and the communicator uses the optimum diversity defined in the Viterbi-Jacobs paper.

In conclusion, what we found is that in the case of FH/MFSK signaling with diversity and partial band noise Figure L-11 we don't have a saddle-point. The communicator should use the value of diversity that was derived by Viterbi and Jacobs, which guarantees a maximum bit-

error rate independent of what the jammer does. Conversely, the jammer should use broadband noise (assuming he's restricted to using noise rather than tone jamming) to ensure a guaranteed minimum bit-error rate. Of course, if possible, both the communicator and the jammer should try to monitor what the other is doing because an adaptive scheme would certainly be able to provide a lot more flexibility than the current system.

FH/MFSK WITH DIVERSITY IN PARTIAL BAND NOISE



REFERENCE:

VITERBI & JACOBS, ADVANCES IN COMMUNICATION SYSTEMS, 1975

B.K. LEVITT PG.1

FIGURE L-1

ASSUMPTIONS

- NEGLECT THERMAL NOISE
- CHIPS ARE INDEPENDENTLY JAMMED, PROB ρ
- PERFECT JAMMING STATE SIDE INFORMATION
- ERROR CAN ONLY OCCUR IF ALL m CHIPS JAMMED
- NONCOHERENT CHIP ENERGY DETECTION
- ALL CHIPS JAMMED ⇒ M-ARY SYMBOL DECISION BASED ON SUBOPTIMUM LINEAR SUM METRIC
- EXACT BER FOR m=1 CASE
- CHERNOFF/UNION UPPERBOUND FOR m>1

PG.2

FIGURE L-2

V-J QUASI-OPTIMUM DIVERSITY

$$\cdot\ \text{BER} \leq \frac{M}{4}\left[\left(\frac{\rho}{1-\lambda^2}\right)e^{-\frac{\rho(\log_2 M)E_b}{m N_0}}\right]^m \equiv f(m,\rho)$$

$$\text{WHERE} \quad 2\lambda = \sqrt{1+6\beta+\beta^2} - (1+\beta)$$

$$\beta \equiv \frac{\rho(\log_2 M)E_b}{2 m N_0}$$

$$\cdot\ \text{BER}\Big|_{m_{opt},\,\rho_{wc}} \leq \min_{m\geq 1}\left[\max_{\rho\leq 1} f(m,\rho)\right]$$

$$= \frac{M}{4}e^{-\frac{\log_2 M}{4}\left(\frac{E_b}{N_0}\right)} \ ; \quad \frac{E_b}{N_0} \geq \frac{4}{\log_2 M}$$

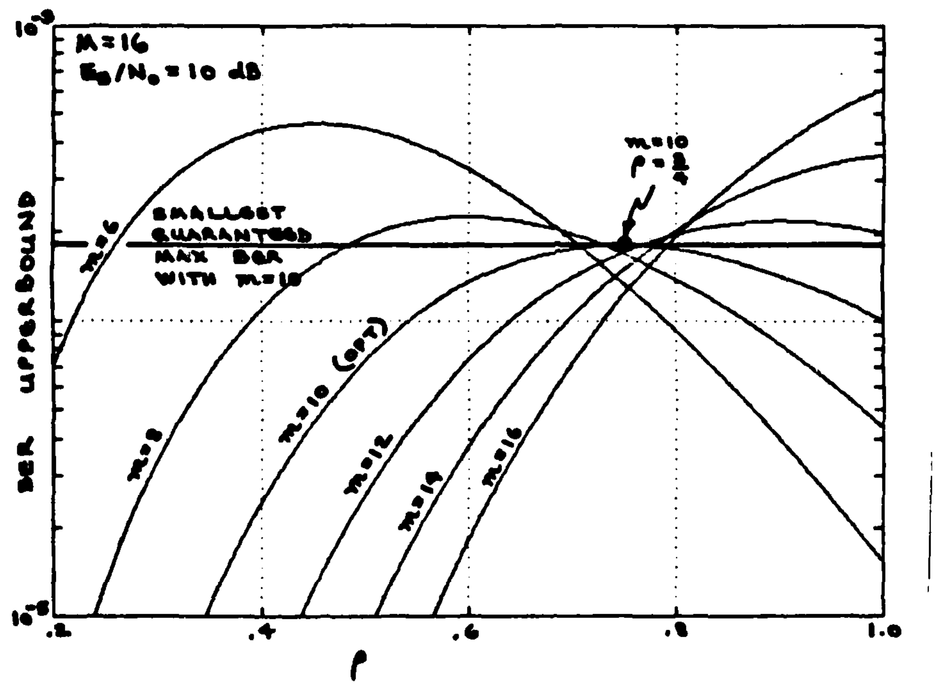
$$\text{WHERE} \quad m_{opt} = \frac{\log_2 M}{4}\left(\frac{E_b}{N_0}\right)$$

$$\rho_{wc} = \frac{3}{4}\ ;\ \forall\, M, \frac{E_b}{N_0}$$

PG. 3

FIGURE L-3

EXAMPLE



PG. 4

FIGURE L-4

IMPLICATIONS

- NO MATTER WHICH m COMMUNICATOR CHOOSES, JAMMER <u>SUBSEQUENTLY</u> SELECTS ρ TO MAXIMIZE BER

$$\Downarrow$$

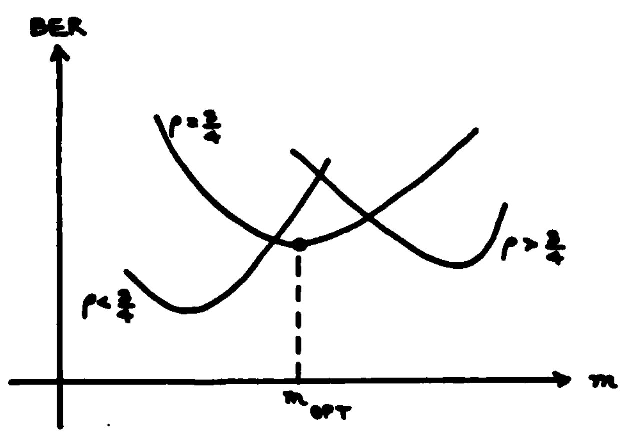
- JAMMER ADVANTAGE
- WORST CASE APPROACH IF ρ CANNOT BE MONITORED
- m_{OPT} ENSURES GUARANTEED MAXIMUM BER

PG.5

FIGURE L-5

IS SOLUTION A SADDLEPOINT?

$$\max_{\rho \le 1} \min_{m \ge 1} f(m,\rho) \overset{?}{=} \min_{m \ge 1} \max_{\rho \le 1} f(m,\rho)$$



- MAX/MIN IMPLIES COMMUNICATOR ADVANTAGE: FOR ANY ρ, m IS CHOSEN TO MINIMIZE BER

FIGURE L-5a

PG.6

REVERSE OPTIMIZATION ORDER

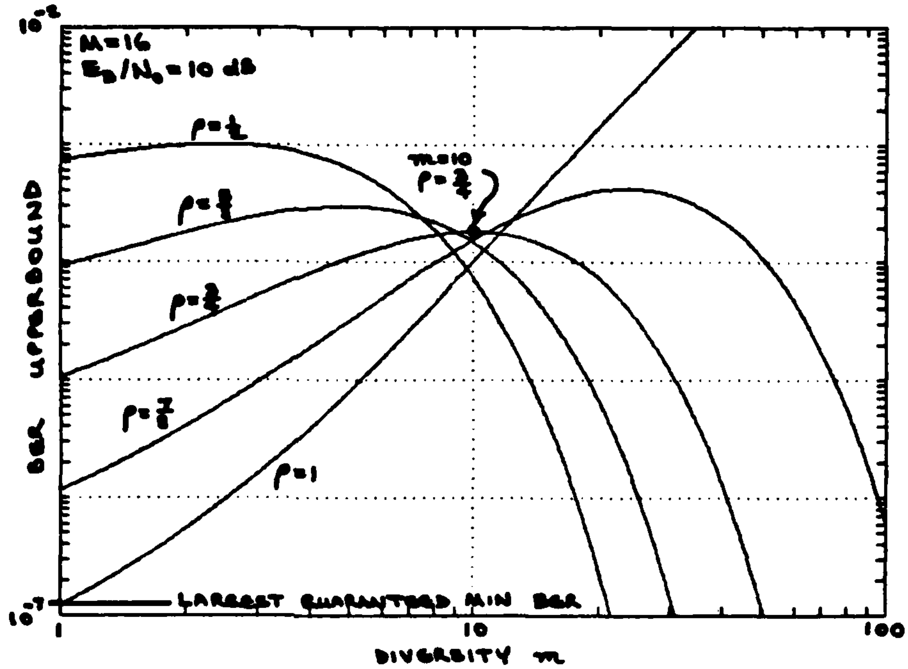


FIGURE L-6

PG. 7

OBSERVATIONS

- BER $\xrightarrow[m\ \text{LARGE}]{}$ 0 FOR ALL $\rho < 1$
- BER INCREASES MONOTONICALLY WITH m FOR $\rho = 1$
- JAMMER CAN ACHIEVE LARGEST MIN BER AT $\rho = 1$: MINIMUM ($m \geq 1$) OCCURS AT $m = 1$
- INTRODUCES GAME THEORETIC CONSIDERATIONS FOR COMMUNICATOR AND JAMMER

FIGURE L-6a

PG. 8

COMPARISON OF MAX/MIN, MIN/MAX

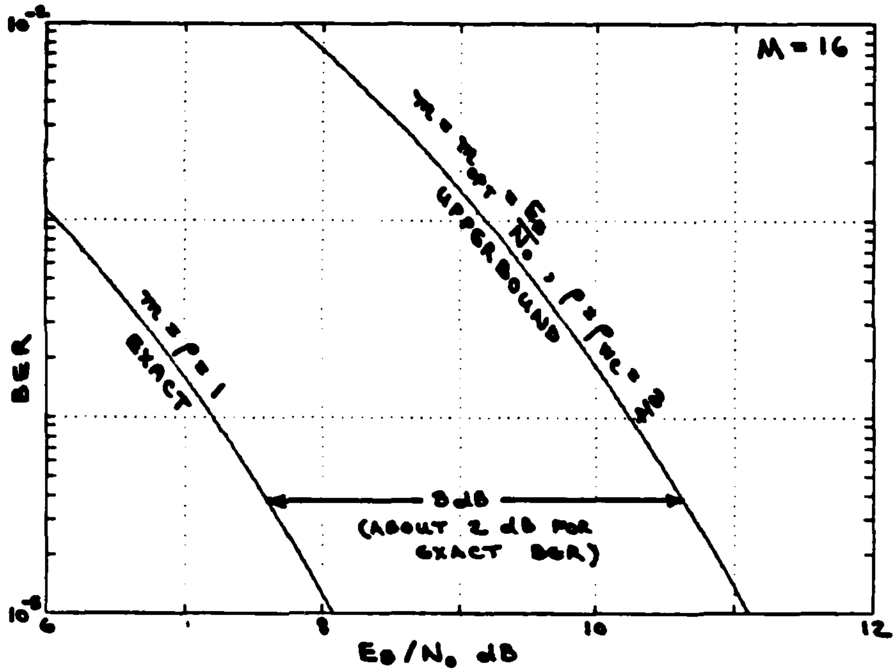


FIGURE L-7

RISK IN USING NO DIVERSITY (m=1)

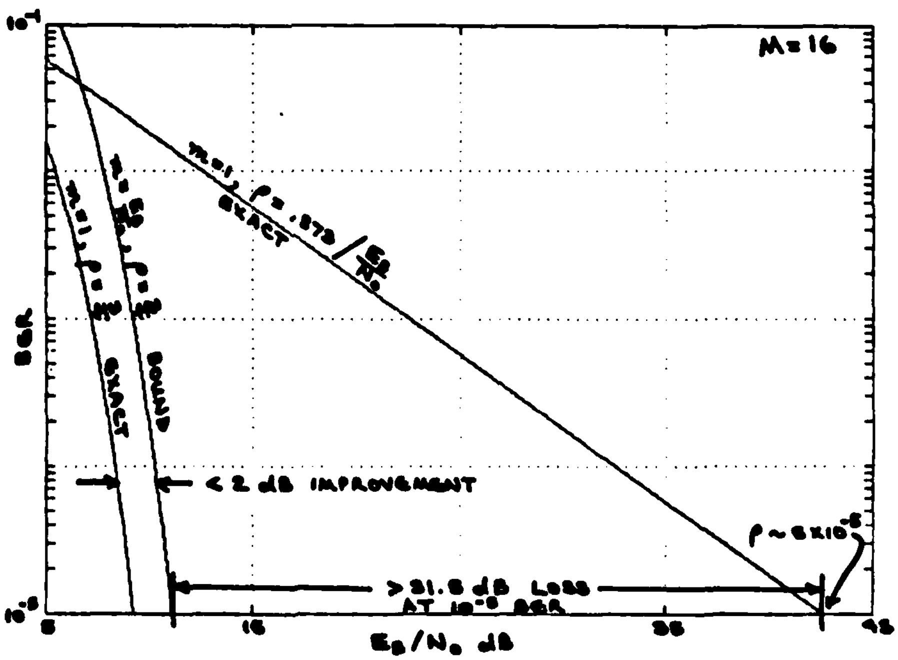


FIGURE L-8

RISK IN USING TOO MUCH DIVERSITY

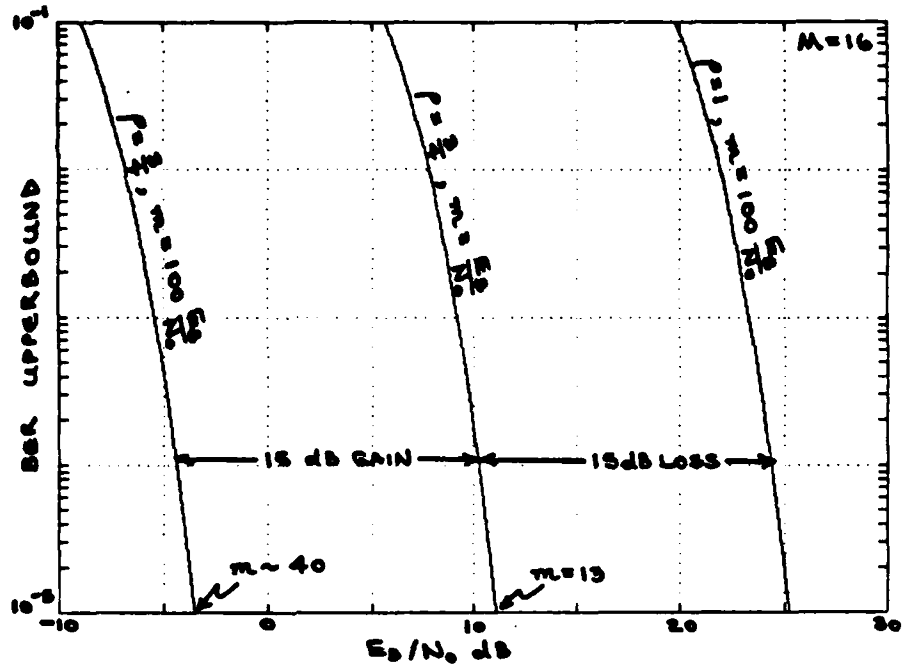


FIGURE L-9

PG. 11

JAMMER PERSPECTIVE : ρ = 3/4 OR 1

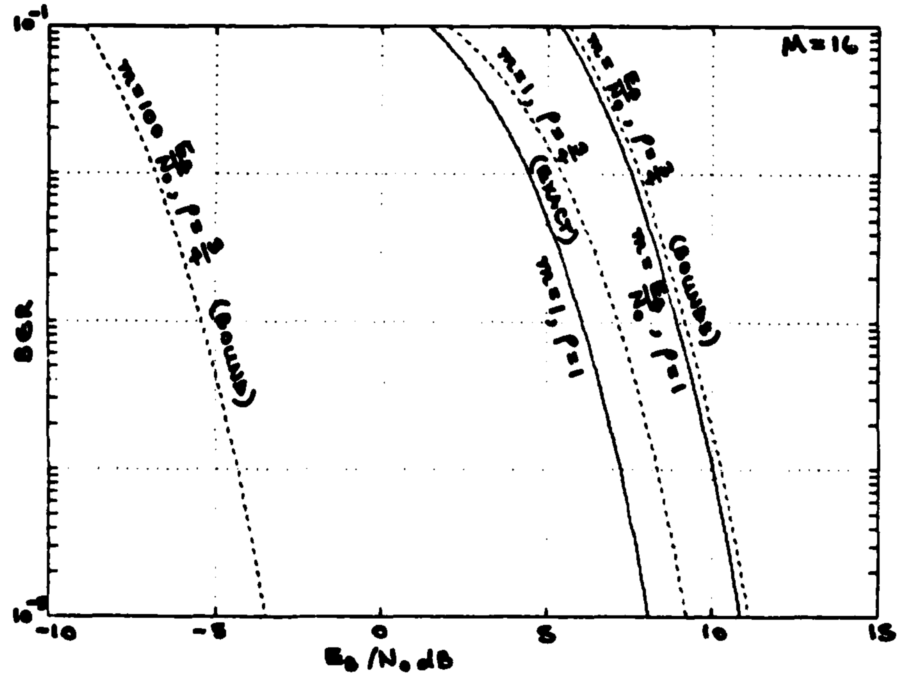


FIGURE L-10

PG. 12

CONCLUSIONS

- $\min_{m} \max_{\rho} \text{BER} \neq \max_{\rho} \min_{m} \text{BER}$
- COMMUNICATOR SHOULD USE $m = \frac{\log_2 M}{4}\left(\frac{E_B}{N_0}\right)$ TO ENSURE GUARANTEED MAX BER
- JAMMER SHOULD USE $\rho = 1$ TO ENSURE GUARANTEED MIN BER
- IF POSSIBLE, COMMUNICATOR AND JAMMER SHOULD MONITOR ρ AND m RESPECTIVELY AND REACT ACCORDINGLY

PG. 13

FIGURE L-11

GAYLORD HUTH

Pravin Jain is the assistant for communication technology in the military satellite communications systems office of the DCA. He received his BS, MS and Ph.D. from the University of Stutgaart, Germany and has been at DCA since 1974. Before that he was at SRI for 6 years.

PRAVIN JAIN

I don't have any esoteric results to show, which doesn't mean that I couldn't have dreamed up some. What I thought I'd do is discuss the role of spread spectrum communications in the design of military satellite communication systems.

The strategic tactical community, as you can see Figure J-1 consists of a diverse mix of user platforms. There is a very substantial population of mobile terminals as you can see here. The data rates however, are fairly modest, ranging from teletype which is 75 bits per second to vocoded voice which is 2400 bits per second. In some cases, users also require what we call tri-tac quality (16 kilobit, CVSD). So these are basically quite modest data rates. However, these data rates have to be supported under stressed conditions. When we talk about stressed conditions, we definitely mean jamming. It turns out that almost all military users require extreme protection against jamming Figure J-2. Selective users also require LPI, low probability of intercept, and some also require protection against nuclear effects. That's one area which is really growing in importance. In the past, a lot of work has been done from the waveform point of view i.e., what signal processing can do to provide resistance against jamming or interception. But really, very little has been done in the past to see how these techniques (which you may want to use for providing protection against jamming or interception) will perform if there are also nuclear scintillation effects. So that area will become more and more important in time. Let me begin with a brief discussion of some jamming threats.

Clearly I cannot talk about jamming threats in a non-classified symposium. However, I'd like to give you some indication from a pure technology point of view what jamming EIRP a jammer might be able to generate at a given frequency. As you know, Figure J-3 EIRP is defined as the product of the power the jammer can generate, and how much antenna gain he can use to focus that power in the direction of the satellite. What I'm talking about is uplink jamming. The jamming threat is determined by the constraint the jammer may have, on how much RF power he can generate at a given frequency, and how much antenna gain he can produce. It turns out that both these parameters are strong functions of frequency. We already have satellite systems in the UHF (300 MHz) and SHF (8 GHz) bands. Then there are some higher bands which we generaly refer to as EHF (30 GHz or 45 GHz) where we might have some military systems operating in the future. At the higher EHF bands fairly large jamming EIRPs are feasible.

What are some of the antijamming techniques available to us? Figure J-4 These techniques we have discussed this morning. Basically we can do spatial processing which we call antenna nulling, or waveform processing. We have two types of waveform processing, frequency hopping and pseudo-noise. We use both these techniques in military systems. Basically frequency hopping is presently used at UHF with mobile type platforms. We heavily use pseudo-noise at X-band. The nulling antenna has been implemented on one of our satellites called DSCS III. So in fact we already use all these techniques in military systems. The next generation of military systems will use onboard processing. In addition to nulling, we

might also do complete despreading in the satellite, possibly also demodulation and decoding. That's why I call it a full onboard processing. Now with a full onboard processing system, the maximum tolerable jammer EIRP can be calculated by the simple formula Figure J-5. If you have a user whose EIRP is S, and the jammer's EIRP is J, then J is usually very much larger than S. The maximum tolerable jammer EIRP is given by this very simple formula, where W/R is the processing gain, W is the bandwidth over which we are spreading, and R is the data rate. Alpha is the antenna nulling, (spatial processing gain) and E_b/N_0 is the modulation efficiency. These are all the parameters we need to evaluate the anti-jam performance.

So let's see with these parameters, what kind of performance we can get at UHF which is about 300 MHz, at X-band (8 GHz), and at EHF. That's shown over here Figure J-6........*the formula* which I showed you earlier. I'm assuming here that the user generates only 10 watts of power at all the three frequencies, and his platform allows him to use a small 2-ft aperture antenna. Now let's look at UHF to see what kind of antijam protection can we provide. The assumption here is that the user wants to communicate at vocoded voice rate which is 2400 bits per second. So it turns out that with 15 dBW of EIRP, all the user can withstand is 53 dBW jamming EIRP without nulling. Now suppose we could do nulling at UHF, and we could pick up 30 dB of nulling, then that number gets up to 83 dBW. That's about the best we can do at UHF.

Now let's see what we can do at X-band. We are still generating 10 watts. Now since we have gone up in frequency, the antenna gain for a 2-footer jumps up from 5 dB to 31 dB. Without nulling all we can withstand is 84 dBW of jamming EIRP. If we were to pick up another 30 dB by antenna nulling, then the protection gets up to 114 dBW. Let's look at EHF. We are generating the same 10 watts and using a 2-foot antenna, so the EIRP is 56 dBW. If we were to spread the signal over the whole available bandwidth, which I'm assuming here is 2 GHz so that means that is 93 dB spreading, with a voice rate of 2400 bits per second which is 33 dB, the processing gain is roughly 60 dB. So without antenna nulling, we get around 105 dBW tolerable jammer EIRP. If we now drop in 30 dB antenna nulling, the tolerable jammer EIRP becomes 135 dBW - a very substantial number.

The point I'm trying to make is that by going to higher frequencies, there's a tremendous payoff. Even a small user terminal is able to withstand a very large jammer. But we don't get this capability by just spreading. We have to do antenna nulling as well as waveform processing. These two must go together if we are really are interested in battling very large jammers

At EHF we are trying to spread the signal over a very wide bandwidth, about 2 GHz. Do we use pseudo-noise or frequency hopping? For mobile type platforms, frequency hopping is a much more attractive approach for easier synchronization and hardware implementation. Another reason for using frequency hopping is superior performance under nuclear effects.

Now let me say a few words about nuclear effects. There's a consensus that soon after the nuclear blast, the natural bandwidth of the medium collapses. It is difficult to say how large then the coherent bandwidth is. It is a function of how big the blasts are, the number of blasts, where they occur and how far away are you from the blasts. The general consensus is that the medium bandwidth shrinks to a fraction of what we had before. So that fraction of the bandwidth

may not be able to support a very wide bandwidth spread spectrum signal. We could in this situation spread the signal only over the narrower bandwidth the medium can support. But if we do that, we lose anti-jam performance. So what do we do to make up the anti-jam? We can take a narrower pseudo-noise signal and then hop it over the entire bandwidth again. Well if we want to do that we might as well use a pure frequency hopping scheme. So that's another reason why frequency hopping may be good.

Figure J-6 illustrates an application to protect low data rates against very large jamming. One point I'd like to make here is that while we are using both waveform and spatial processing, the most anti-jam performance is coming from waveform processing. There's about 60 dB at EHF from waveform processing and 30 dB from nulling. So most of the anti-jam performance is coming from processing gain. Traditionally when we talk about military satellite communication, we are talking low data rates. For low data rate applications pseudo-noise and frequency hopping are the two areas where most of the research and effort has been devoted. We did not pay much attention to antenna nulling until we started looking at systems at X-band like DSCS III. Now let's look at a different situation.

Suppose we have the following scenario Figure J-7. We have a satellite, two terminals, a transmitter and a receiver, and we want to have very high data rate communications between the two terminals. I am assuming 100 megabits, at an uplink frequency of 30 GHz. So this is an example of high data rate communication at EHF. There's a big jammer present too. This is a very different case than what I considered earlier. The earlier was a small platform with a low data rate against strong jamming. This new example is the case of a big terminal trying to transmit a very high data rate, 100 Mbps against jamming Now let's see what kind of performance we get here Figure J-8. Let's say the user transmitter can generate 1 kilowatt of power, so that is 30 dBW and say he's using a 40 ft antenna at 30 GHz, a gain of 68 dB. This means that the user can generate roughly 98 dBW of EIRP. Now we have allocated 1 GHz of bandwidth at 30 GHz so if the data rate is 100 megabits, then the processing gain is only 10 dB. Remember in the case considered earlier we were spreading over 2 GHz at a data rate of 2400 bps so the processing gain was roughly 60 dB Here, the processing gain is only 10 dB which is very low because the data rate is very high. So after going through this very simple calculation we find that the maximum jammer EIRP which the user can support is only 98 dBW at 100 megabits If we want to withstand roughly the same jamming level as in the case of small mobile terminals then we would require 40 dB of spatial processing. That would bring up the maximum tolerable jammer to 138 dBW. Here we have the same situation, we are doing both spatial processing as well as waveform processing. But here now the situation is reversed. Here we are getting very little from waveform processing (spread spectrum) but we are getting a lot from antenna nulling Now this is the area which is not very well understood In the military community, people are very jittery about nulling. You have to know how far away the jammer is with respect to you. Because if you null the jammer, you null the user also, if he is located very close to the jammer. That's an area where work has to be done.

One other point I'd like to make. I didn't say anything about coding, and I know there are a lot of coding theorists here I am assuming an E_b/N_0 of 10 dB Anything we can do with coding to lower E_b/N_0, will lower the amount of antenna

nulling that is required. With Viterbi decoding, or some other decoding of your choice, how much coding gain can we get? Let's say we pick up coding gain of 5 dB so the required antenna nulling now becomes 35 dB, which is still a large amount of nulling. As far as spread spectrum processing is concerned, we are only talking about 10 dB of waveform processing. That may be something we might be able to do with pseudo-noise, with today's high speed digital circuitry, and high speed logic. There may be a possibility to take the 100 megabit signal and spread over 1 GHz, using pseudo noise and pick up 10 dB processing gain. That doesn't mean that we may not be able to do that with frequency hopping also. But if we want to do frequency hopping, then what is the hopping rate we are going to pick? If we pick a low hopping rate compared to the data rate (200 khps or something like that) we have to transmit many bits of symbols per hop. To optimize performance with frequency hopping, we may have to do interleaving. Pseudo-noise for this kind of application may be a better scheme from a pure implementation viewpoint.

In summary I would like to say that when I look at the gamit of applications we have in military communications, I find that there is room for both pseudo-noise and frequency hopping. We really have to look at the application. We can't say that one is always better than the other. There are applications for one and applications for the other. For small mobile-type terminals, low data, very low EIRP, trying to fight large jammers, I think frequency hopping might be a good way to go. There is however, one problem. Since users have different data rates, each data rate leads to a different hopping rate for maximum jamming. So if we want to have a common transmission format where everyone hops at a common rate, then some may lose in jamming performance. So here we have a problem that does not exist in the pseudo-noise area. Another reason for using frequency hopping might be for nuclear effects. Large bandwidth spreading might not be workable in a nuclear environment. However for high data rate communications, I think pseudo-noise starts becoming very attractive, as I mentioned earlier, anything we can do in the coding area to minimize E_b/N_0 will immediately help in the antenna nulling area. I very strongly urge that we communication theorists who have spent a lot of time working in the waveform processing, should now start looking in the antenna nulling area. That's a fascinating field but it's fairly virgin so far. Only antenna-theorists have dabbled in that area. I think there's a big challenge there. Start looking into it, because really, we are going to ask for high data rate communication for imagery transmission and things like that, and we are not going to get very much protection from waveform processing. Clearly, the antenna-nulling area - the spatial processing - is going to be very, very important.

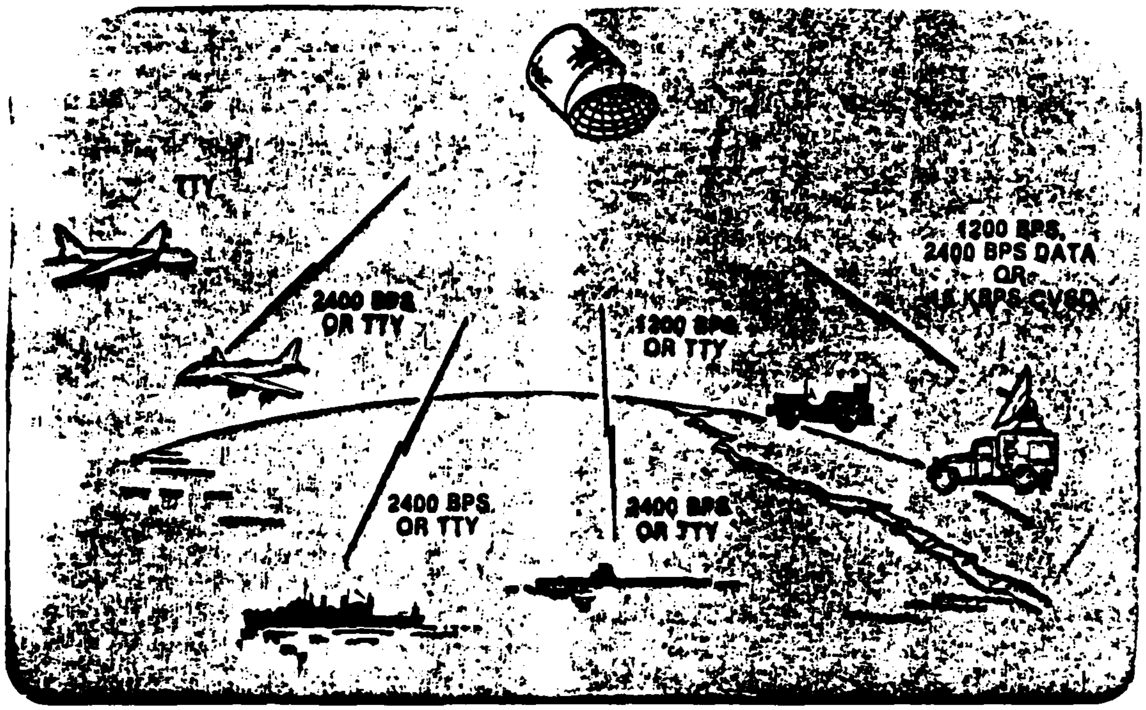


Figure J-1 STRATEGIC/TACTICAL USERS

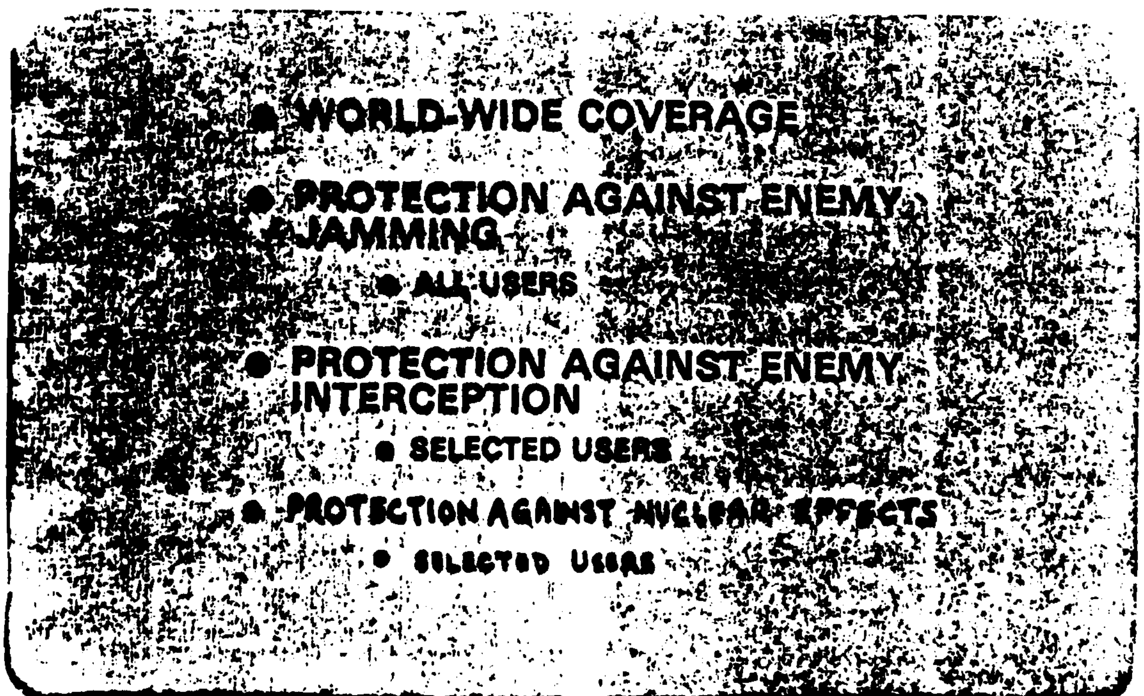


Figure J-2 STRATEGIC/TACTICAL USER NEEDS

JAMMER EIRP (dBW) = JAMMER POWER (dBW) + JAMMER ANTENNA GAIN (dB)

- MAXIMUM JAMMER EIRP IS CONSTRAINED BY:
  - JAMMER'S POWER GENERATION CAPABILITY
  - JAMMER'S ANTENNA DIAMETER AND ANTENNA POINTING ACCURACY

Figure J-3 JAMMING THREAT

- NULLING ANTENNA
- SPREAD SPECTRUM
  - FREQUENCY HOPPING
  - DIRECT SEQUENCE PSEUDO NOISE

Figure J-4 ANTI-JAM TECHNIQUES

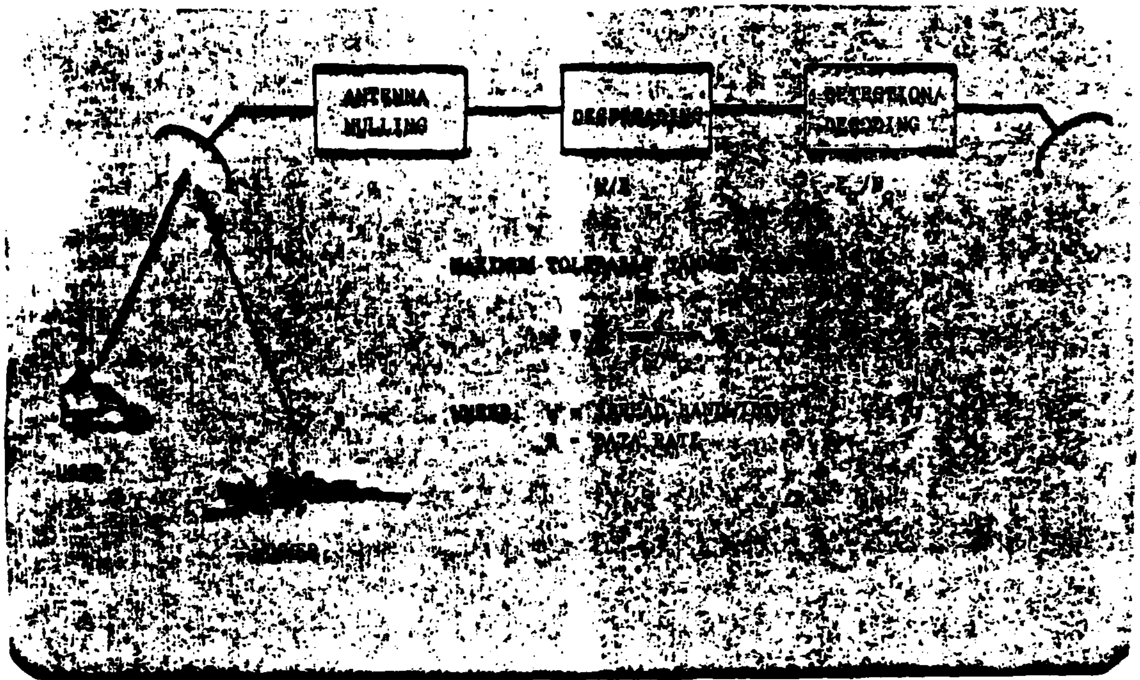


Figure J-5 ANTI-JAM PERFORMANCE (ONBOARD PROCESSING)

ANTIJAM PERFORMANCE

		UHF	SHF	EHF
TRANSMITTER POWER	10 W	10.00 dBW	10.00 dBW	10.00 dBW
ANTENNA APERTURE	2 FT.	5.00 dB	31.00 dB	46.00 dB
TRANSMITTER EIRP		15.00 dBW	41.00 dBW	56.00 dBW
SPREAD BANDWIDTH		82.00 dB (160 MHz)	87.00 dB (500 MHz)	93.00 dB (2000 MHz)
DATA RATE	2400 BPS	33.80 dB	33.80 dB	33.80 dB
E_b/N_o		10.00 dB	10.00 dB	10.00 dB
MAXIMUM TOLERABLE JAMMER EIRP		53.20 dBW (3 KW/10 FT)	84.20 dBW (5 KW/15 FT)	105.20 dBW (10 KW/20 FT)
ANTENNA NULLING		30.00 dB	30.00 dB	30.00 dB
MAXIMUM TOLERABLE JAMMER EIRP WITH NULLING		83.20 dBW (200 KW/35 FT)	114.20 dBW (400 KW/45 FT)	135.20 dBW ??????

Figure J-6

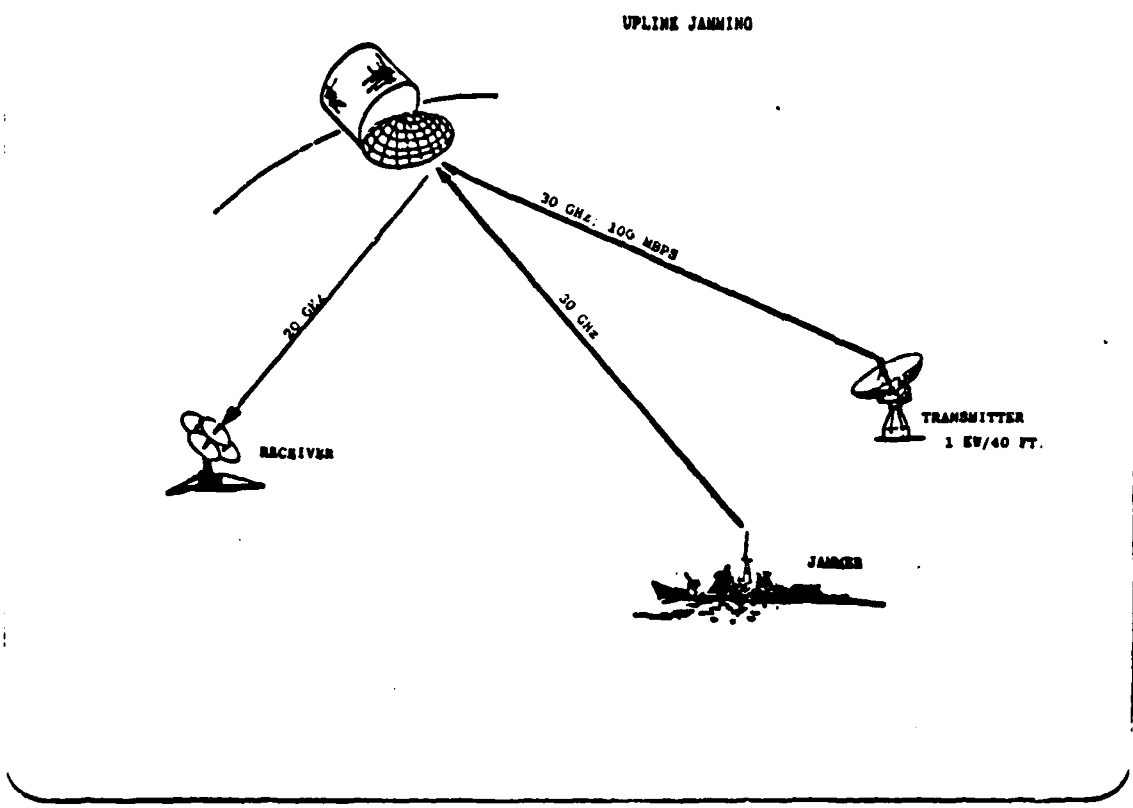


Figure J-7 HIGH DATA RATE COMMUNICATIONS

ANTI-JAM PERFORMANCE

TRANSMITTER POWER	1 KW	30 dBW
ANTENNA GAIN	40 FT/30 GHz	68 dB
TRANSMITTER EIRP		98 dBW
SPREAD BANDWIDTH	1 GHz	90 dB
DATA RATE	100 MBPS	80 dB
PROCESSING GAIN		10 dB
E_b/N_o	PSK	10 dB
MAX TOLERABLE JAMMER EIRP		98 dBW
ANTENNA NULLING		40 dB
MAX TOLERABLE JAMMER EIRP WITH NULLING		138 dBW

Figure J-8

GAYLORD HUTH

Jerry Gobien is at DARPA in the Tactical Technology Office. He received his Ph.D. from the University of Michigan in 1973. He taught communication and radar at the EE Department of the Air Force Institute of Technology in Dayton, Ohio from 1973-1977. From 1977 to 1981, he was the manager of Advanced Development of Air Force's Tactical Jam Resistant Voice Communication Systems - SEEK-TALK. Since 1981 he's been at DARPA-TTO mainly working with radars and sensors, but he maintains a spiritual commitment to communication systems

JERRY GOBIEN

I stand before you with considerably reduced innocence; about 5 years ago, I was unceremoniously removed from my college teaching job and put into one where I was asked to actually develop one of the systems we've been discussing. Not just a breadboard kind of thing, but a system that was actually to be fielded by the Air Force The job was complete with all of the political battles, funding hassles, and challenges that go with a program where the "end of the rainbow" is a billion dollars in production money I wanted to come here to tell you of the things we did right, those we did wrong, and, for those who were familiar with the program, to tell you whatever happened to SEEK-TALK.

The problem that I will be addressing is that of jam-resistant tactical communications. This morning, Marlin Ristenbatt added the word "mobile" in there and I think that's appropriate. Figure G-1 It implies an environment where everything is constantly moving; you haven't a chance of "pointing at" the terminal that you are trying to communicate with because typically you don't know where it is or even what it is behind. The locations are unknown. The transmitter in this situation can't usually be relied upon to add order to the network. When the war starts, the "transmitter" is typically scared, or at least under a lot of stress, and apt to do some irrational things The challenge is to develop a "protocol" that places order upon this network.

I assert two things Figure G-1: (1) In this communications situation, you cannot come up with the right technical solution until you thoroughly define what I call the "networking" requirements and I will talk about those in just a minute The problem with these requirements is that they are very difficult to define because doing so involves a lot of work on the part of the customer for the system. He, on the other hand, is not apt to put an awful lot of work into defining such requirements until he knows that the system is "real", which means that the technology must be well-defined and at least partially proven. Clearly, there is a kind of tail-chasing to this. My second assertion is that (2) I don't think that we have the technology in hand to meet the Air Force's requirement for jam-resistant tactical communications under the constraints that you believe the user's stated requirement, and that you believe allocated RF frequencies are needed in order to ultimately operate the system The Army's problem is an order of magnitude more difficult and, by implication, we therefore cannot solve that one either with today's technology

Let me talk a little bit about networking requirements of Figure G-2 I'm going to assume (and for the Air Force problem it's a realistic assumption), that the required data rates are known and fixed Under "network requirements", I don't include data rates and queueing times (the things that digital communication people normally think of as network requirements). I really mean the more fundamental business of defining and describing "nets" The Air Force situation

seems easy at first blush; you just define them by analogy to radio channel. The tactical air communication system today uses a UHF AM radio, and users who currently communicate on one frequency channel constitute a "net" in a jam-resistant system. You then need think a little bit more about how many jam-resistant nets you need and how to establish an intra-net protocol which preserves the current "free-wheeling" structure of a radio channel. You need to worry both about the number and the geographic distribution of nets and of users within a net. Furthermore, it matters how many users in a net are apt to be transmitting at the same time. The scenario I described is really a broadcast scenario, and the number of people pushing their microphone buttons at the same time becomes a real issue. It implies an interesting question at each receiver: "Of all the users simultaneously transmitting, how many are required to be received at the same time?" (I'll talk about that more specifically in a moment.) Finally, if that number is greater than 1, how are you going to select them?

But let's assume you have done all that and you believe everything you've come up with down to here Figure G-2.(1). You will still have a very large problem: You positively will not get any more frequencies than are currently allocated to military communications. Based upon hard experience, I will guarantee that it's not going to happen. You may get an experimental allocation in a microwave band somewhere, but that's about it. Moreover, the existing bands already have users in them. These people tend to jealously guard the allocations that they have, and they are neither going to go away, stop communicating, nor take it very kindly if you interfere with their current operations.

So you end up with a fairly complicated net structure once you actually start defining this system, and it's one that dramatically affects your choice of technology (i.e., whether you use frequency hopping, direct sequence, or hybrid spread spectrum, how much antenna nulling do you need, should you use directive transmit antennas, etc. etc.) The interaction between the "network" requirements just listed and the performance of a particular choice of technology is really quite dramatic. For instance, you are going to be jammed by existing users of the band. They may be lower in power, but it is quite a shock to discover that they are not all low power. For example, we discovered quite late in the SEEK-TALK program that there are European-manned UHF communication sites (part of the NATO air defense system) which, knowing that they will be jammed, already have the problem solved. They have "bootleg" 1 to 5 kilowatt power amplifiers in their basement and, when they get jammed, they are going to "burn through". There are quite a few such sites in Europe; the upshot is that all it really takes to jam the UHF band is one enemy jammer turning on some place, causing all of these guys to turn on their kilowatt power amplifiers, and we will have UHF gridlock. In addition to existing users, you also have to worry about yourself. This becomes a large issue in deciding what waveform to choose. It is not much of a problem for a frequency hopper that's narrowband instantaneously, but becomes serious for direct sequence spreading with overlapping frequency bands. In that case, your own transmitters become tremendous partial or full-band jammers for you.

The third implication wasn't really appreciated until we started considering AWACS in detail. The AWACS has a requirement to operate on many nets simultaneously (they like to think that they're controlling all those airplanes out there). There is a phenomenon that occurs

on the AWACS which is even now poorly understood. It's suspected to be a "rusty bolt" effect which occurs where various plates on the fuselage touch each other. There's corrosion and oxide in between, and this has the same effect as distributing small diodes on the skin. When you have two simultaneous transmissions the third and fifth order intermodulation products caused by these nonlinearities can fall back into the original frequency band. The power level of these intermodulation products can be as high as that of a transmitted signal originating 50-60 miles away All of the above effects combine to reduce the "available" jam resistance of the system.

Another thing to worry about is the effect of your new system on current users of the band. Those "other guys" are typically narrowband radios. Their receivers all have squelch circuits, which keep them from making a lot of noise (amplified receiver noise) unless they are *actually receiving something*. To upset the existing system, you need only "break" those squelch circuits; this will produce a tremendous hissing noise, the operator will think he is being jammed, and he will raise holy hell with you. This effect occurs at much lower power levels than are *required* to *actually "jam"* his *radio* so he couldn't receive a transmission But as soon as he hears that receiver go off and start making great amounts of noise you've interfered with him.

The list of "net requirements" and their implications is not complete, nor is it optimally thought out. But I assert, probably trivially, that you've got to take them all into account when choosing a technical solution. If you do, you will end up with a different answer than if you had done a 1 on 1, or even a single net or small-number-of nets analysis.

Enough generalities. Let me talk about a case study, the SEEK-TALK system. SEEK-TALK was a program to develop an Air Force jam resistant voice communication system. It was to be an improvement or replacement for the UHF AM radio which, as every communicator know, is quite vulnerable to jamming. The "need" was surfaced by events of the 1973 Arab-Israeli war, coupled with the critical dependence of tactical air operations on voice communications among pilots and between pilots and the TACS The program was kicked off by a study which was intended to consider all of the things on the previous slide In fact, it did so better than is done for most development programs, but still not well enough to hold together in the end The study was convened, and the requirement was looked at in some detail and defined The different technologies available to satisfy that requirement were examined. The candidates are shown on the slide Figure G-3,(1). All frequency hopping approaches were rejected, and direct sequence spread spectrum with adaptive antenna arrays was chosed. With 20-20 hindsight, *I think* the last two were rejected correctly even given today's technology. The fast frequency hop/direct sequence hybrid was rejected for reasons which probably were erroneous. (Lincoln Labs, upset at the decision, went ahead and constructed a proof by example.)

Let me elaborate a moment on the original requirements Figure G-4. Some of these were explicitly stated in the study, and some were really just assumed. The first two are pretty straightforward, the key one that was really just "assumed" is the third. We never made a big issue of it and the effect later in the program was dramatic. Our guess at number of nets, though not contradicted by the user, wasn't advertised very well by us Figure G-4,(1). It was based on a scenario where a whole gaggle of airplanes is heading toward combat, all talking to each other on the way in. That is considered

important because pilots will listen to the radio traffic and form a mental picture of what's going on, where the threats are, who is being shot at, where the command control platforms are, etc. Then, when they actually engage in combat, they typically break off and work on a smaller network with a forward air controller or with their own flight. As soon as the engagement is over, they switch back to the large net in order to rejoin the broader activity and keep track of the action. So the assumption was that we needed 5 or 6 nets; a couple of small ones and one "common strike" net that people would come in and go out on. We estimated 50 users per net; that's probably high for the small nets and low for the common strike net, but it's a reasonable average. About 10 of these users would be transmitting simultaneously. The next requirement was among the most interesting and controversial: Each receiver has to be able to receive a number of transmissions simultaneously, and those ought to be selected on the basis of range from the receiver (closest being the most important). The idea is that when a man in your own flight yells "Break left!", you want to be sure you heard that (presumably he just saw somebody about to shoot at you). If somebody 20 miles away says something, that's not as important. Furthermore, a break-in capability isn't good enough. Pilots are quite adept at listening to 2 or 3 things at the same time and picking up the important key words. This came to be known as "conferencing" and it was a requirement which I think still stands today. There were ground and airborne command and control stations that have to service all users simultaneously from a great distance. Acquisition of transmissions had to be done instantaneously (instantaneously means 2/10th of a second in voice communications; if it is slower, your ears perceive the delay). The structure was to overlay the current tactical air control system, and hence was pretty well constrained.

The technical solution Figure G-5 after examining all candidates was direct sequence spread spectrum, roughly 5 megachips per second spreading of 16 kilobits CVSD voice. That was the lowest voice data rate that we felt could use in a combat environment with high background noise. Within each net, simultaneously transmitted signals had different spreading codes, chosen adaptively (one can listen, see which codes are being used, and then choose a different one when the mike button is pushed). Rubidinm time standards permitted range measurement for the purpose of selecting the closest transmissions. DS Spread Spectrum alone provided in sufficient jam resistance; we needed another 25 dB of nulling from an adaptive antenna to get the needed JR without high powered amplifiers. (We already knew that frequency management would be a problem.)

And so we launched a development program Figure G-3,(2). The process was intense in every aspect, from technical three political to personal. There seems to be a never-ending series of crises and challenges to the program, for everything from budget to technical reasons, from NSA doesn't like your spreading codes to the fellow who wants more business to the one who has a better solution. At the same time, and this was probably fatal to SEEK-TALK, your cost estimates grow because you get a better look at what the hardware is really like At each program challenge, the user (TAC) perceives that their program, (it is now <u>their</u> program, they've developed some advocacy for it) is in trouble, so they try to help by strengthening the requirements. This is a subtle and key point. As the user support and "need" for the system grows, the

"network requirements" subtly grow along with it. In SEEK-TALK's case, they grew (almost unnoticed, at first) to the point where they were technically unsupportable.

This programmatic merry-go-round continued for four years, sweetened somewhat by a successful test of the advanced development hardware. It was stopped by the arrival of a new administration which observed that if all the Air Force's on-going tactical communications programs (JTIDS, SEEK-TALK, and Mark-15 IFF) ran to completion, the cost would be nearly $15 billion. SEEK-TALK itself had grown from an initial estimate of $900 million to $3 billion. Since that much money was out of the question, an intensive review of the on-going programs was directed. The issues were: Interoperability between services, cost, jam resistance and robustness, and potential commonality of functions The Air Force Chief Scientist chaired this study. We had the user (TAC) heavily involved in refining and reviewing the operational requirement. I participated with the jam-resistance panel in trying to evaluate all the systems against a common threat. For the first time, the frequency management people participated heavily; some were brought over from Europe to reveal what was in those bands and what the effect of inserting SEEK-TALK would be (This was very late in the program to be doing such analyses.) It was a long and protracted study and I think a fairly sound one. The results are summarized in Figure G-7. Astoundingly, the operational panel concluded that we needed in excess of 200 nets in a 100x100 km area (The structure we had assumed within a net was valid.) No viable technical solutions (even the ones we had rejected) will support that requirement in the UHF band, especially if you take the existing users of the band into account. If you ignore these EMI and EMC issues, then both the Direct Sequence/Adaptive Array hybrid and the FH/DS hybrid were viable candidates. The fast frequency hop/direct sequence spread system is JTIDS-inoperable in the following sense: Although JTIDS as it currently exists does not have the required jam resistance, it is possible to invent a mode with more direct sequence processing gain to yield the needed jam resistance. Such interoperability is not possible using a direct sequence system like SEEK-TALK but is needed because, using it, you could talk in a jam-resistant mode to the Navy. It was this consideration which finally caused the SEEK-TALK technical approach to be abandoned in favor of the FH/DS hybrid system which is now called HAVE-CLEAR.

I won't dwell on the points of this slide Figure G-8; they are the recapitulation of my rambling of the last half hour. You need to work on all those things (networking, jam-resistance, and spectrum requirements and availability) together. It is very important, if you really expect to deploy a JR system, that you understand the frequency allocations process and its politics. Otherwise, you might naively assume that some day you might get a piece of a radar band to communicate in (or a radio astronomy band, or almost anything else.) If I had to vote for where we should put technical effort, it's in the areas that allow you to operate in the face of all these real-world problems rather than in trying to figure out how to build a device that can process signals with 20% relative bandwidths.

JAM-RESISTANT TACTICAL COMMUNICATION

RE:

- o *FLUID, HIGH-STRESS ENVIRONMENT*
- o RECEIVER OFTEN CAN'T BE DESIGNATED
- o STATIONS ARE MOBILE
- o LOCATIONS TYPICALLY UNKNOWN

1. TECHNICAL SOLUTION CAN'T BE CHOSEN UNTIL "NETWORKING" REQUIREMENTS KNOWN IN DETAIL. THESE ARE VERY HARD TO DEFINE UNTIL THE TECHNICAL SOLUTION DEMONSTRATES SOME SUCCESS...

2. THERE IS NO DEVELOPED TECHNOLOGY TO SOLVE THE AF'S REQUIREMENT UNDER THE CONSTRAINTS OF REQUIRED NETWORK STRUCTURE AND AVAILABLE FREQUENCY BANDS. (THE ARMY"S PROBLEM IS MUCH HARDER...)

FIGURE G-1

TYPICAL "NETWORK REQUIREMENTS":

- DEFINITION OF "NET"
- NUMBER OF NETS
- GEOGRAPHIC DISTRIBUTION (NETS AND USERS)
- NUMBER USERS/NET
- NUMBER OF SIMULTANEOUS TRANSMISSIONS
- NUMBER OF SIMULTANEOUS TRANSMISSIONS TO BE RECEIVED
- SELECTION OF CRITERIA

MILITARY COMM. BANDS ARE WELL KNOWN AND FIXED. THEY WILL NOT MOVE OR GROW, NOW WILL THEIR CURRENT OCCUPANTS GO AWAY (OR EVEN KEEP QUIET)

IMPLICATIONS

- JAMMING BY OTHER "FRIENDLY" USERS
- SELF JAMMING AND NEAR/FAR SUPPRESSION OF MULTIPLE D.S. SIGNALS
- INTERMOD. PRODUCTS (COSITE)
  REDUCED AVAILABLE J.R. (NOISE AND 'SMART')
- EFFECT ON OTHER USERS...(SQUELCH!)

TAKING THESE INTO ACCOUNT WILL RESULT IN A SIGNIFICANTLY DIFFERENT ANSWER THAN A '1 ON 1' OR EVEN 'SINGLE NET' ANALYSIS.

FIGURE G-2

CASE STUDY

AF'S J.R. TACTICAL VOICE

1974-5 o NEED WAS IDENTIFIED: REQUIREMENT STUDIED AND DEFINED.

1976 o TECHNICAL APPROACH WAS SELECTED (DS + AA, FFH/DS, FFH, SFH)

1977 o DEVELOPMENT PROGRAM STARTED: SFH NEAR TERM, DS/AA LONG TERM

1978 o RECURRING CHALENGES TO PROGRAM

- o COST ESTIMATES GREW

o REQUIREMENTS ('NETWORKING') GREW

1981 o SUCCESSFUL ADM TESTS

FIGURE G-3

ORIGINAL REQUIREMENTS

(SOME EXPLICITLY STATED, SOME ASSUMED...)

- JR: 1:5 RANGE DISADVANTAGE AGAINST A NUMBER OF HIGH-POWERED JAMMERS, POSSIBLY RESPONSIVE.
- FREQUENCY: UHF (225-400 MHz)
- 1 LARGE AND 4 OR 5 SMALL NETS IN A COMBAT AREA, CLOSE IN TO JAMMERS.
- 50 USERS PER NET, 10 SIMULTANEOUSLY, SELECTED IAW RANGE FROM RECEIVER ("CLOSE IS IMPORTANT").
- GROUND AND AIRBORNE C^2 STATIONS SERVICE ALL NETS SIMULTANEOUSLY.
- "INSTANTANEOUS" ACQUISITION
- VOICE QUALITY PERMITTING RECOGNITION
- ESSENTIALLY AN OVERLAY TO THE CURRENT TACS: REQUIREMENTS FORMULATED BY ANALOGY TO CURRENT SYSTEM AND ANALYSIS OF SEA TAPES.

FIGURE G-4

ORIGINAL SOLUTION
"SEEK TALK"

- DS (~ 5 Mc/s) SPREADING OF 16KB/s CVSD, INTRA-NET CDMA.

- ATOMIC TIME STANDARD PERMITS RANGE ORDERING.

- ADAPTIVE NULLING ANTENNA ARRAY ON RECEIVE: NEED 25 DB AGAINST SEVERAL JAMMERS.

ADVANTAGES

- o REQUIRES NO HPAs.
- o LIMITED BW HELPS FREQUENT MANAGEMENT, FORCES JAMMER SIGINT.

DISADVANTAGES

- o ARRAY VULNERABLE TO SIGINT DIRECTION OF LARGE NUMBER OF JAMMERS.
- o SIGNAL MASKING EFFECTS.

FIGURE G-5

CASE STUDY

1980-1981

- o NEW ADMINISTRATION CHARTERS 'COMPREHENSIVE REVIEW' OF ALL 3: VOICE, DATA, IFF.

ISSUES:

- INTEROPERABILITY BETWEEN SERVICES
- COST AND POTENTIAL COMMONALITY
- JR AND ROBUSTNESS

PANELS:

- OPERATIONAL..."WILL THE TRUE REQUIREMENT PLASE STAND UP..."
- JAM RESISTANCE..."AGAINST A COMMON BASELINE THREAT, AS WELL AS RESPONSIVE OPTIONS. MANY ON MANY.
- EMC AND FREQUENT MANAGEMENT..."WHAT IS THE ENVIRONMENT REALLY LIKE? WHAT ARE THE RESULTING PROBLEMS?
- IFF..."IS JTIDS USEFUL? IS THERE SOME USEFUL FEATURE OF THE OTHER CONTENDERS?"

FIGURE G-6

WORKING GROUP RESULTS

1. THE "TRUE REQUIREMENT" EXCEEDS 200 NETS IN A 100 × 100 KM AREA.
2. THE ASSUMED INTRA-NET REQUIREMENTS WERE VALID.
3. THE UHF BAND IS A HORRIBLE EMC ENVIRONMENT, UNPREDICTABLE COUNTRY-BY-COUNTRY.
4. THE ONLY VIABLE TECHNICAL SOLUTIONS ARE, INDEED, DS + AA AND FFH DS.
5. NEITHER ONE CAN SUPPORT THE REQUIRED NUMBER OF NETS IN THE MILITARY UHF BAND.
   - FFH AND SFH COULD, BUT SUFFER OTHER PROBLEMS
6. BOTH SUFFER ULTIMATE VULNERABILITIES:
   - DS/AA: SIGINT DIRECTED TO OVERCONSTRAIN AA.
   - FFH/DS: PROLIFERATION OF "DUMB" UNATTENDED PARTIAL BAND JS.
7. SEEK TALK IS UNIQUE AND OFFERS NO HOPE FOR INTEROPERATBILITY. FFH/DS CAN BE STRUCTURED TO INTEROPERATE WITH JTIDS (NAVY) VOICE, THOUGH JTIDS ALONE HAS INSUFFICIENT JR.

FIGURE G-7

REPRISE

- o NETWORK STRUCTURE/SIZE, SYSTEM JR, AND SPECTRUM REQUIREMENT ARE CLOSELY RELATED.
- o THEY MUST BE UNDERSTOOD, AND COMPROMISES MADE, AS PART OF THE PROCESS OF CHOOSING A TECHNICAL APPROACH.
- o THE FREQUENCY ALLOCATION AND MANAGEMENT PROCESS IS ILL-UNDERSTOOD BY TECHNICIANS.
  - THAT'S BECAUSE IT'S INCREDIBLY ILLOGICAL AND POLITICAL.
  - IT'S ALSO OF CRITICAL IMPORTANCE.
- o WE MUST DEVELOP TECHNIQUES TO PROVIDE JR IN THE FACE OF THESE REAL-WORLD CONSTRAINTS:
  - LIMITED BW, LARGE NUMBER USERS
  - SELF AND CO INTERFERENCE
  - IM PRODUCTS, OUT OF BAND RADIATION
- o HAVING DONE THE BEST WE CAN, SOME COMPROMISE OR REQUIREMENTS WILL BE NECESSARY. THIS IS EASIER AT THE BEGINNING THAN 3-4 YEARS INTO A PROGRAM.

FIGURE G-8

GAYLORD HUTH

Seymour Stein is co-author of a couple of comm. books He started in spread spectrum in the mid 60's, including tropo-scatter in 1963. I personally know of a 1965 project that he led at Sylvania, because that's what got me into spread spectrum. They published a classified 800-page compendium of spread spectrum theory at that time which told me where to go, and it was quite good. The last 4 years he's been an independent consultant operating under the name SCPE, Inc.

SEYMOUR STEIN

I have the unenviable job of following 4 well-qualified speakers. As somebody said, we were also uncoordinated, so I don't think I'm going to cover any one topic that hasn't been covered by someone else, but I certainly don't agree with everything that's been said. I guess that's what a workshop is all about.

There have always been two ideas about what spread spectrum is for Figure S-1. One is the anti-jam role which we've been talking about. The other is the LPI role. LPI in the past has almost always been described in terms of covert operations, the idea that somebody really does not want anyone else to know that he's broadcasting. I'm beginning to get a view that LPI in another form is going to become significantly more important in the near future, and that is not to deny the fact that you're broadcasting, but to try to avoid being targeted (physical destruction). Some useful systems now being designed may not avoid detection, but may be able to avoid being targeted. I think that's an important idea to keep in mind in designing future systems.

A second thing which I think you've heard in spades already is that in the spread spectrum world, we really do not have one problem, we have many problems Figure S-2. We sometimes forget that there are different problems and we try to come up with all encompassing solutions. Right now, I can identify three different sets of problems I know about. One is the HF world where there's an awful lot that can be done in the electronic warfare sense, to take advantage of the characteristics of propagation at HF, but about which unfortunately very little is being done On the other hand it's also a very difficult world in which to work. It's probably even more difficult than the UHF-VHF world in which to work spread spectrum, because the bandwidth is really not there. In the VHF-UHF world, I think Jerry (Gobien) has said it all, but I'm going to make some comments about networks which maybe diametrically opposite to some of Jerry's conclusions about where the future may lie. Then finally there's the radio relay world, including satellite relay There the difference is that we have directional rather than non-directional antennas, very wide bandwidths are typically available because we're up in the microwave bands and the data rates can vary all over the map depending upon whether it is a sole user or a high level network that is being supported.

In the next chart S-3, I tried to recreate for this meeting some data from many years ago. Back in the middle 60's, one problem that we had to contend with was a lot of military people who kept saying, "What's with this spread spectrum? You're asking me to give up megahertz of bandwidth and all I get is a single voice channel?" We had to convince them on channel capacity arguments that the voice channel was all they had available under poor SNR conditions. Well, when I redid this, I came up with a peculiar result which I don't recall in the earlier interpretations. I translated the channel capacity into a required processing gain to

get another version of the chart that Pravin (Jain) was talking about earlier. The peculiar result that pops up says that you'd better not forget that in addition to the jammer there is receiver noise. If I read this correctly, it says that Pravin's statement to the contrary, if the coding people try too hard to achieve a system that will work at a very low E_b/N_0, they're going to cause trouble in terms of the required processing gain to cope with jamming. Finally, I've put down at the bottom, a different version of why it's so hard to do the problem by processing gain, namely I've tested some bandwidths that are typical of what people talk about, and some very modest data rates. As Pravin showed us, it's very easy to generate threats that indicate some of those numbers are simply not enough.

In the last few years, my main interest in this area has been in the tactical comm band, the one Jerry Gobien has been describing Figure S-4. I think this is a particularly nasty problem, just as he said, but have perhaps summarized it in a different way. The traditional frequency management problems are just terrible. The way in which frequencies are long-term reserved by all kinds of users out there, independent of actual usage, simply makes life difficult for other people, and the system makes it politically impossible to argue with them. My own feeling is that technical work is somehow needed in this area to plan a long term future in which everybody recognizes that frequency is a resource and that somehow it's got to be allocated almost adaptively on the basis of need or demand.

A second comment deals with the LPI aspect Figure S-5. We keep running into confusion when we talk about jammer strategies. In the tactical world we've had a tremendous amount of clamor about what the frequency-follower-jammer can do to communications, but even if we resolve that, how the enemy will simply go into barrage jamming. I want to present on this chart, at least one version of the story. It starts out with one interesting assertion, which some people believe. That is that nobody can really afford to go about randomly jamming the entire band, because the enemy also has C^3 requirements. What that means is that they do have to jam responsively. (By the way, this is obviously a good example of something that might apply in the mobile tactical world, but that might not apply to jamming an up-link on a satellite. There it can be pretty clear whose satellite you're jamming.) One interesting consequence is to really think about how to defeat a responsive jammer, one that tries to follow you. Perhaps the best way is to simply make sure that you don't have one system or one net, or even a few nets out there using the same type of modulation. Get as many people out there operating in the same mode as possible, even to the point of using decoys if necessary. Another thing you can do, which is not done, is to think about a realistic approach to power level management to further reduce detectability. Then if you want to really get fancy, start varying the level that you use just to make it difficult for someone to spotlight you. Finally, a tendency in the past in designing equipment has been to always go for the best that technology can provide. One result is that we now produce equipments that they can be distinguished one from another by very small parameter differences. It appears that we may be a lot better off if we quit insisting on very stringent specifications in some of the parameters, just to foil easy recognition.

There has been much talk about direct sequence operation or even hybrid operation with direct sequence Figure S-6. Whenever anybody says hybrid, they are still talking about an instantaneous bandwidth of the order of 1 MHz, 5 MHz,

20 MHz, depending on what band is in mind. But in this tactical world, you simply cannot co-exist with the narrowband users; you cannot live with them until something better is developed that exists now in the way of notched filters, spectral whitening or adaptive interference cancellation, so that can get rid of those that are too near to your receiver. What can be done in the way of modulation is not an issue any longer. That is not the problem. The problem is to come up with workable and reasonable-cost equipment that can get rid of the narrowband interference which otherwise represents non-hostile jamming that can be worse than anything the enemy can throw at you.

I will also say that earlier when Barry Levitt was talking, a comment flitted through my mind. I think the famous phrase going something like this, "Those who do not read history are condemned to repeat it." What went through my mind is that there's been a great deal of concern in the past few years about the inadequacy of frequency hopping in the face of rather sophisticated receivers that are quite capable of rapidly following any kind of hopper. In contrast, when I was first introduced to frequency hopping, nobody ever thought that you would get by with just hopping a carrier. The thought was that you would have to devise a data system that avoids any susceptibility to frequency following. Figure S-7. It has recently been reinvented, and is now called independent frequency hopping. As an example, if you want to use 8-ary FSK, you select for each symbol 8 independent tones that the receiver would listen for, one of which would be transmitted by the transmitter. An enemy could hear the tone transmitted but would not know what the other 7 are and have no idea how to find them. So he couldn't do anything other than a pure random guess. The reason that independent hopping fell out of favor, particularly at data rates like 2400 bits per second, was simply technology and cost. I think the time has come when the technology is now just about caught back up. Synthesizers are getting cheaper, and there may still be a need for clever ways of designing receivers to use them to make the cost even lower. To repeat something that Ristenbatt said this morning, one may have to worry not so much about demodulation as how to achieve rapid initial sync in the mobile tactical environment. That can be a formidable problem when you have to acquire so-called code time. There are devices that could solve the problem, devices that basically implement a filter bank instead of implementing a number of independently tuned filters. They are not being adequately exploited at this time for spread spectrum, and for the life of me, I don't know why.

Now, I'm ready to try to pick up where Jerry Gobien left off. The solution to the ground tactical environment is not nulling because you can't build complex enough antennas on a mobile platform to take care of the jamming environment. It is also not to try to get wide bandwidths in the lower UHF band because the spectrum availability is just not there. You have to go way up in frequency. Jerry said microwave, and may be right, but you may not be able to get radar users to move over and allow you to operate at 7 or 8 GHz, so maybe you have the ability to design small antennas with narrow beams, and narrow beams are very nice to have in an EW environment. If you're lucky, the jammer won't hear you and even if he hears the transmitter, he may not be able to get in on your receiver. Furthermore, if you give up on the notion that everybody has to hear everybody else (omni-azimuthal environment), and go to networking, you can avoid the problem instead of trying to brute force a solution. As far as I know, relatively little is being done in this direction. Anyone looking at it

will recognize lots of issues here that require plenty of 6.3 money, 6.2 money, whatever you want to call it. It's a research program tactical communication for the 90's that's a very different approach - any kind of network is a very different approach than the mobile tactical user is used to. He's used to having that little whip antenna out there, so that anybody who is supposed to hear him is going to hear him, yet I don't see how we can live with the electronic warfare threat that keeps growing every year until we start taking advantage of some avoidance possibilities.

This is the min-max problem in another version Figure 9. A week ago, at the Comm Theory Workshop, Joe Aein tried to provoke a heated discussion by saying that modulation and synchronization theory are dead. By which he meant that we better quite beating a dead horse and look at some of the other problems. When he said that, what registered in the back of my mind was that it certainly applied to "spread spectrum modulation all over again." The earliest spread spectrum systems were invented to deal with the WWII class of jammers, a simple transmitter that was tuned to a frequency that you were using. The earliest responses, the earliest anti-jam or LPI systems, were very simple, and they were designed to cope with this idiot narrowband jammer. Sure enough, it didn't take very long before people recognized that the jammer could also get smarter. If the jammer knows you're hopping (we talked about super-het receivers) he can follow you, and maybe you can't hop fast enough to keep out of his way. That's the reason for independent hopping. But then you find that if you're using too small a set of frequencies, he can engage in partial band-jamming that can be very successful against a simple hopper. This potential clever jammer, jamming just a few tones of our frequency hop set, can cause an unreasonable error rate and leave us with practically no A-J capability. Again, the answer was very simple. The earliest frequency hop system was fielded with a (15,5) triple error correcting code. Very simple, very effective, by any kind of calculation. Moreover, if the jammer concentrates his resources in any one part of the spectrum or any one part of the time-domain, you don't let that energy come roaring through. You clip it or put a limit on it before you use it in demodulation or in decoding. At that point you very quickly discover that the potentially clever jammer has now been negated to the point where not only have you removed the susceptibilities, but by the fact that he was trying to attack them, and that you've coped with that attack effectively, he almost doesn't do anything at all to you anymore. That quickly says, OK, what's left for him is a more robust jammer operating at the min-max or max-min point. The problem is that I think we've been there for almost 15 years in the modulation area, and that there's little more to do. Moreover, I think Dave Chase and others have pointed out the road for the ultimate spread spectrum communication design. It is to regard our resource as a time-frequency domain with degrees of freedom that greatly exceed the information requirement. Then, instead of spread spectrum, think about how to generate and use a very low rate code in that domain that can operate successfully within the very poor signal-to-jammer ratio, and the spectral coloration of the jamming. I think that is precisely what has really been considered in many of the studies of spread spectrum modulation, but too indirectly.

Of course, another question is, what is the ultimate jammer design? He's not a constant envelope jammer. He's someone who is noise-like. And, despite my earlier disbelief about barrage jamming, that

certainly would be the ultimate jammer. He has all of the random time and frequency variations that you would have with white noise and fully occupies whatever bandwidth and time domain are available. Having recognized all that, I suspect that we are very close, in a dB sense, to the best we can do, and I think we do have to look for other techniques.

We always talk about the frequency domain Figure S-10. From time to time, people raise the question of time hopping. Sometimes we study it. Some day we probably will have good amplifiers where we can maintain a high average transmitted energy even while operating on a low duty cycle, and then time hopping will join the rest of the resources we have. It's not really going to resolve the limitations of the frequency-time environment for lots of netted users, but it has advantages. It requires no new analysis. It's really just another variation on the same theme Finally, if we look at the minimax solution, if we say that we will force the enemy towards as much barrage jamming as he can afford as his best strategy, are we really being realistic? I don't know. Since we also have to design our own jammers, maybe there is a role here for some research to understand how one can be jamming and communicating at the same time. I don't know of any such research. Maybe there's a way of doing it if you constrain the deployments of the jammers. I think that it may be an interesting topic. But if the answer to this is really null, then regardless of some of my earlier negative comments about life in the UHF band, maybe there are ways to make life uncomfortable enough for the frequency follower jammer so that he really will have to give up on jamming by following, but also won't be able to jam us simply by barraging the whole band with all his energy. If so, what then are the real minimax or maximin strategies for us, the communicators?

ROLE OF SPREAD SPECTRUM — AND ITS LIMITS

ANTI- JAM : LIMITED BY PROCESSING GAIN AVAILABLE

LPI :
- AVOID JAMMER SET-ON
- AVOID DETECTION OF OPERATION (COVERTNESS)
- AVOID BEING TARGETED

FIGURE S-1

DIFFERENT APPLICATIONS POSE DIFFERENT PROBLEMS

LONG-HAUL HF (INCLUDING LF-MF)	STRATEGIC C^3	NON-DIRECTIONAL ANTENNAS MANY MOBILE USERS HEAVILY CONGESTED BAND MODEST INSTANTANEOUS BANDWIDTH

PROPAGATION PATH AVAILABILITY A MAJOR FACTOR (BOTH SIDES)

NETWORK/RELAY POSSIBILITIES (CURRENTLY POINT-TO-POINT)

HF — UHF	TACTICAL C^3	NON-DIRECTIONAL ANTENNAS MOBILE USERS /NETS HEAVILY ALLOCATED BANDS (POSSIBLY EXCEPT L-BAND) MODEST-TO-LARGE INSTANTANEOUS BANDWIDTH VOICE AND DATA REQUIREMENTS

LOS JAMMING < GROUND VS. AIRCRAFT / AIRCRAFT VS GROUND > NUMBERS ADD TO SEVERE THREATS

NETWORKS WILL REPRESENT A TOTALLY NEW CONCEPT

RADIO RELAY VHF - EHF (INCLUDING SAT. RELAY)	TELECOMMUNICATION NETWORKS STRATEGIC, CIVIL, HIGH ECHELON TACTICAL	DIRECTIONAL ANTENNAS (JAMMING VIA SIDELOBES, EXCEPT SATELLITE UPLINKS) VERY WIDE BANDWIDTHS AVAILABLE DATA RATES - SINGLE CHANNEL TELEGRAPHY TO MULTICHANNEL TELEPHONY

FIGURE S-2

PROCESSING GAIN ALONE CANNOT HANDLE THE LONG-TERM JAMMING ~~PROBLEM~~ THREAT

$$R \le W \log_2\left(1+\frac{S}{N+J}\right) = 1.44\, W \ln\left(1+\frac{S}{N+J}\right) \rightarrow 1.44\, W \frac{S}{J+N} \quad b/s$$

$R = 1/T \qquad T = \text{AVER. BIT DURATION}$

$$1.44\,TW \ge \frac{J}{S} + \frac{N}{S} = \frac{J}{S} + \frac{n_0}{E_b} TW$$

$$TW \ge \frac{J/S}{1.44 - 1/(E_b/n_0)}$$

(NOTE THAT CUDDLING UP TO SHANNON LIMIT AGAINST ADDITIVE NOISE IMPOSES GREATER PROCESSING GAIN REQUIREMENT AGAINST JAMMING)

EXAMPLES:

W	R	TW
10 MHz	50 b/s	53 dB
100	"	63 dB
1 GHz	"	73 dB
100 MHz	2400 b/s	46 dB
1 GHz	"	56 dB

FIGURE S-3

PROBLEMS IN CONVENTIONAL TACTICAL COMM BANDS

(MOBILE USERS --- HF-UHF BANDS)

- PROCESSING GAIN LIMITED BY BANDWIDTH TOTALLY AVAILABLE
- REAL ESTATE LIMITS { DIRECTIONAL BEAMS / DEGREES OF FREEDOM IN NULL-STEERING
- SPREAD SPECTRUM MUST COEXIST WITH CONVENTIONAL NARROW BAND USERS
- TRADITIONAL FREQUENCY MANAGEMENT PROCEDURES ARE DEADLY TO SPREAD SPECTRUM
  - LONG TERM RESERVATION APPROACH MAKES WIDEBAND ALLOCATIONS ARGUMENTATIVE
  - FIXED ALLOCATIONS (EVEN IF WIDEBAND) SPOTLIGHT THE SPREAD SPECTRUM NETS
  - NEED APPROACHES THAT ALLOW "ADAPTIVE" ASSIGNMENT

FIGURE S-4

IMPORTANCE OF CONFUSION IN THE TACTICAL ENVIRONMENT
(INTERMEDIATE SOLUTION)

- ASSUMPTIONS: NEITHER SIDE CAN AFFORD INDISCRIMINATE WIDEBAND ("BARRAGE") JAMMING BECAUSE IT NEEDS ITS OWN C^3
  IN ANY CASE, HAMPER TARGETING
- HOW TO SOW CONFUSION:
  - WIDESPREAD USE OF SAME TYPE SPREAD SPECTRUM SHARING SAME BAND
  - USE OF DECOY TRANSMITTERS
  - TRANSMITTED POWER LEVEL MANAGEMENT (MINIMUM TO REDUCE DETECTABILITY, VARYING FOR CONFUSION)
  - REDUCTION OF INDIVIDUALIZED EMITTER FEATURES
- RESULTS OF CONFUSION
  AJ: JAMMING INTERMITTENT, CODING PULLS TEXT THROUGH
  ANTI-TARGETING: INCONSISTENT FIX FOR TARGETING OR HOMING

FIGURE S-5

NARROWBAND INTERFERENCE EXCISION – REQUIRED FOR D.S. OPERATION IN CONGESTED BANDS

- IMPLICIT IN FREQUENCY HOPPING (USE OF CODING TO OBVIATE HITS)

TOO COSTLY FOR MODEMS?

- SPECTRAL ANALYSIS WITH CONTROLLABLE NOTCH FILTERS
  - RESTRICTED TO BANDWIDTHS ALLOWING DIG. SIGNAL PROCESSING?
  - POSSIBLE ROLE FOR RAPIDLY-PROGRAMMABLE SAW DEVICES?
  - NOTE NEED FOR SIGNAL DELAYS TO DO THOROUGH JOB IN DYNAMIC (PUSH-TO-TALK) ENVIRONMENT
- ADAPTIVE SPECTRAL WHITENING
  - WIENER FILTERS
  - LPC
- ADAPTIVE INTERFERENCE CANCELLATION (INTERFERENCE IDENTIFIABLE INDEPENDENT OF SIGNAL)
  - WIDROW ITERATIVE ALGORITHMS
  - CAN COMBINE WITH CHANNEL EQUALIZATION

→ NOTE IMPLEMENTATION ISSUES (TECHNOLOGY) MAY BE FREQUENCY-DEPENDENT

FIGURE S-6

SIGNIFICANT ROLE FOR INEXPENSIVE FREQUENCY SYNTHESIZER

- OR BETTER WAYS TO USE SYNTHESIZERS
- OR LESS EXPENSIVE EQUIVALENTS

- INDEPENDENT FREQUENCY HOPPING (ONE OR MORE HOPS PER SYMBOL) NEGATES THREAT OF FOLLOWER JAMMER
  - MULTIPLE HOPS PER SYMBOL MAY APPROACH ULTIMATE LPI

- M-ARY FSK RECEIVER FOR IFH REQUIRES M SYNTHESIZERS
- RAPID INITIAL SYNC ACQUISITION WITH N-HOP COMBINER REQUIRES M×N SYNTHESIZERS

- POSSIBLE EQUIVALENT: WIDEBAND SPECTRUM ANALYZER (FILTER BANK)
  - DIGITAL (NOT IN SIGHT AT LOW COST FOR EVEN 1MHz)
  - SAW RAC DEVICES?
  - BRAGG CELL DEVICES?

FIGURE S-7

MY GUESS AT LONG-TERM FUTURE TACTICAL COMMUNICATIONS

(MANY RESEARCH ISSUES)

- TRANSCEIVERS AS NODES IN HIGHLY FLEXIBLE, MULTIPLY-CONNECTED NETWORK

- MOVE TO MICROWAVE OR MILLIMETER BAND
  - NARROW BEAMS
  - WIDE SPREAD-SPECTRUM BANDWIDTHS
  - ABOVE 15GHz, ABSORPTION VS. RANGE

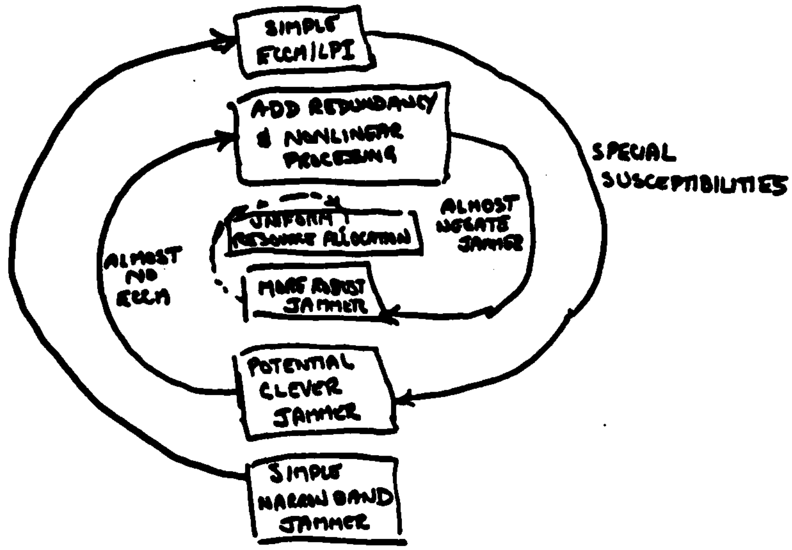
- HIGHLY ADAPTIVE CONNECTIVITY/ROUTING
  - RAPIDLY STEERABLE BEAMS (PHASED ARRAYS OR RAPIDLY SWITCHABLE ASSEMBLIES)
  - SOUNDING/SPATIAL SEARCH TO ACCOMODATE RAPIDLY MOVING PLATFORMS (AIRCRAFT)
  - SINGLE TRANSCEIVER TIME-SHARED OVER ALL LINKS PER NODE
  - RAPID REROUTING/RELAY RESPONSE TO JAMMING

- VOICE ENTRY WITH AUTOMATIC RECOGNITION, FOR LOW DATA RATES

FIGURE S-8

NEED TO REEXAMINE WHERE WE'RE AT IN THE MODULATION/CODING (SPREAD SPECTRUM) GAME

ARE WE ALREADY NEAR THE MINIMAX/MAXMIN ?



ULTIMATE COMMUN. DESIGN (?) - LOW RATE CODE
ULTIMATE JAMMER DESIGN (?) - WIDEBAND, NOISELIKE (STOCHASTIC TIME & FREQUENCY VARIATIONS)

} 100% F-T OCCUPANCY

FIGURE S-9

SOME FINAL COMMENTS

- MUST RECOGNIZE SEVERAL DIFFERENT ENVIRONMENTS (E.G. GROUND TACTICAL VS SAT. RELAY)
  - DIFFERENCES: SPECTRAL BACKGROUND
    ANTENNA DIRECTIONALITY
    PROPAGATION, DELAY EFFECTS
- NULLING IN REAL ENVIRONMENT WILL HAVE TO COPE WITH HIGHLY DYNAMIC SIGNAL CHANGES
- POSSIBLE FUTURE FOR TIME-HOPPING
  STILL AWAITING LOW DUTY CYCLE, CONST. AVG POWER AMPLIFIERS
- IS BARRAGE JAMMING REALISTIC IN THE TACTICAL ENVIRONMENT?
  - CAN COMMUN. BE COORDINATED WITH JAMMING SO ONE'S OWN C^3 SNEAKS THROUGH?
  - WILL IT REQUIRE SPECIAL JAMMER DEPLOYMENTS?

FIGURE S-10

SPREAD SPECTRUM COMMUNICATION IN JAMMING

DISCUSSION

WELCH to GOBEIN: What is the general experience with combining adaptive arrays with spread spectrum on the SEEK-TALK program?

GOBIEN: This was planted, but not at my instigation. We struggled long and hard during the program with the question of how to try to interface adaptive arrays and the spread spectrum modem. If you remember back in the time frame, the chart showed that this system was being designed about 1977. If you think there's little known about how to do that today, you should have been there in 1977. There were speculations about the whole range of things that you could do. But we never felt terribly comfortable with the same thing that people this morning didn't feel comfortable with. That was, "How would the modem synchronize in the presence of this adaptive array trying to hop on every piece of energy that came in from some direction in space?" So we backed away from the kind of feedback systems where the modem would acquire signal and feed that signal back to the adaptive array and keep the array from nulling signals that had the proper spread spectrum code on them. We never felt comfortable that those systems would actually synchronize in the long run. So what we settled on as a structure was the half-way in between system, and I thought it was fairly clever. I have to give credit to Hazeltine. They are the people that actually thought of this. Hazeltine in their system used a very very fast surface acoustic wave matched filter. Fast in the sense that it could be reprogrammed with a new reference code for all intents and purposes instantaneously. So with a couple of SAW devices, they could do a number of things. It was fast enough that we could afford to put a matched filter on an omni-directional antenna element and a matched filter on the adaptive array output. We had a circuit in there that would determine which of those two devices had at any instant of time a better signal-to-noise ratio and switch the signal processor to that spigot. So if we got into a situation where the strongest thing out there was in fact the signal that you wanted to listen to, the array, that was unconstrained by knowledge of signal structure, would null that signal. But on the other hand, that signal would come across just fine in the top channel and we would still process it. If the signals were weak relative to the jammers then this thing would process the bottom channel. There were situations in between, especially in the face of a large number of jammers (when the array was almost fully constrained or over-constrained), when that selection circuit would get a little schizophrenic and flaky, and hop back and forth between various things and never could quite make up its mind. My guess is that had the program been pursued to a next phase, that was a hardware-software problem which could have been worked out but we never tried it.

What we did is build this thing and the array was anywhere from 5 to 7 elements. We had 7 elements on the airplane, we would use 5, 6 or 7. It was space tapered in the sense that in the 7-element configuration, 4 of those were actually half wavelength spaced. The remaining elements were farther out in a position that, when fully constrained, would give you some grading lobes or grading nulls. On the other hand in an under-constrained situation, it would give you much sharper nulls than you would get with an array where all the elements were half wavelength spaced. The nulling

algorithm was a simple analog, least-mean squared, Howells-Applebaum configuration. We didn't know how to do sample matrix inversion rapidly enough at that point to make us feel comfortable if we wanted to try it. So that's what it looked like. What we did was build these things. That makes it look very simple, this thing was a rack that high and that wide, packed full of about 4 million dollars worth of hardware, and it was a truly sophisticated machine, made sophisticated largely by this requirement for conferencing, instantaneously acquiring, processing the n closest, and if a new signal came up, dropping the farthest away. Getting it to do all that in real time was pretty tough. We tested these things extensively first in a laboratory ANECHOIC chamber, in an upside down captive airplane that was sitting on a tower on a mountain that we could rotate and illuminate with various jammers, and finally on a T-39 that we flew down at Eglan. The thing that *absolutely* blew my mind was that it really worked and it worked quite well. We achieved the jam resistance that we were looking for. I think the array rather consistently (when it was not fully constrained or over-constrained) achieved the 25 dB of cancellation that we were looking for, against 2, 3 or 4 jammers. We really hadn't expected that.

We were worried about masking effects, what happens when a signal is aligned spatially with the jammer. In the ANECHOIC chamber with nothing moving, everything calibrated, and sitting still, we got masking over roughly a 5 degree slice around the jammer signal. If you snuck the signal in that closely, two and a half degrees on either side, you couldn't acquire it anymore. Once we got away from a flat ground plane and put this thing on an upside down A-10 on a tower, that grew to about 10 degrees for a number of reasons, primarily multipath and polarization effects I think, because the elements are now sticking out at different angles. In flight test, as best as we could tell from reduced data, it was maybe a 15 degree slice, and that's getting borderline. If you can't hear a signal plus or minus seven and a half degrees from a jammer, at long distances, that is a big hunk of real estate in which you cannot see signals.

This was one of the things that worried me about the results of the flight test. The other one was that a lot of pattern simulation had led us to think that the arrays would degrade gracefully as they were over-constrained. We were set up to simulate and to actually do a very sophisticated simulation of the flight test scenarios before we went down and flew them at Eglan. We simulated multipath, the actual jammer layout, and the signal layout. Then we tried to run the airplane through it and see where you could communicate. That simulation led us to believe that the degradation would not be nearly as graceful as we had hoped. The *flight test* results in fact bore that out. So I think if there was one result from all of this that made me truly queasy, it was that with a 5-element array, when you have 6 jammers out there, all of which are of roughly equal power, you had a real problem. That in turn impacted our assessment of how serious the SIGINT directed threat was. The assumption was that if the guy had to put 12 jammers on the net to really kill you, and find the net first, maybe that wasn't so serious On the other hand, if you only had to put 5 on there, then you had more problems. So I think of all the test results that we got, this was the one that worried me the most. This was the one that I was equally concerned about before hand and even the simple-minded array turned out to be surprisingly robust to blinking jammers, frequency switching jammers, all of the standard kind of responsive threats that you can think of. We paid NRL a whole bunch of money to play to red team to us,

and they put together a fairly sophisticated jamming system where they tried blinking, switching, coordinating and all those standard things. We had beaten the contractors pretty severely with that threat throughput the development program so they all had carefully designed fast attack, slow decay AGC circuit. That turned out to be, at least with all of the things we tested, very little of a problem. The one that concerned me on the airplane was that one.

WELCH to LEVITT: In the game, you did have a saddle point. In the classical game theory when that happens, you usually imbed the game in a larger game, like putting a probability distribution on the parameter space. In looking at this particular game, putting the ρ in a probability space wouldn't buy you anything, but having the diversity M random would. Did you look into that and see what the value of the gain was?

LEVITT: It seems like randomizing the diversity M certainly is providing an additional degree of freedom. It will probably buy you something. I am not sure exactly what it will buy. There are systems being built that have variable A-J capabilities. The whole issue of strategies about reacting to changes in an environment to control these parameters is being studied.

McELIECE: It is still not obvious that there's going to be a saddle point because of the convexity properties of whatever this payoff function is. But having random diversities certainly looks like something interesting. I think there is likely to be a saddle-point. They key is whether it acts in the randomization coefficients.

POSNER: It seems to me that it probably does have a value. The key thing is the dependence on the randomization coefficients rather than on the communication theory stuff that enters into the error-probability. So I'll give you 3 to 1 that it has a value.

WELCH to STEIN: This is a sort of a comment. I forgot who was talking about or suggesting that the solution to the tactical problem was to go up to millimeter waves. My question is what do you do about pointing in a tactical situation?

STEIN: The concept was one that started with a development at Lincoln Lab which the people at MITRE had taken note of. It was a cylindrical, rapidly steerable array. The idea would be to take this one array and time share it over many wings emanating from a node very rapidly. If you look at any one node, it would be rapidly rotating both in a continuous search mode to acquire new people, and also to convey in a packet sense, information very rapidly when you needed to get it. The attraction to that kind of thing is noted on my chart. It begins to introduce a lot of additional elements into the problem of how to minimize the effect of the jammer. Namely it gives you lots of possibilities for alternate routing. I personally think it has to be a deterministic kind of thing only because in any application like that, the data rates would be so high that only a deterministic control system could handle it. Nevertheless, it would have adaptation features.

PURSLEY: It seems to me that in this game theory formulation, one of the factors is the cost to the jammer of the fractional band that he must occupy. Not only in terms of the equipment that he has to build, but also, as has been pointed out, it precludes his use of that same channel as well. The other point is that perhaps our analysis should not be over-reliant on the Gaussian distribution, because, in fact, the jammer may not be very Gaussian-looking. We took as a goal for a given bit error rate making the percentage band the jammer has to jam in order to have any

possibility of reducing the bit error rate below that, regardless of what type of jammer you use. For example, for 85% of the band, a 32-12 Reed-Solomon code with 6-order diversity accomplishes that. The jammer has to hit more than 85% of the band in order to have any possibility of getting you below a bit error rate of 10^{-3}. This is with 32-ary orthogonal modulation. The nice thing about that is you can apply that result to multi-user communication, where you have the self-interference, hostile jamming, or whatever.

COMMENT-HUTH: I'd like to comment, and it goes along with what Pravin was saying earlier. We've all done a lot of research in the area of anti-jam, but the problem is that if you look at these cost numbers, and you look at how big that jammer gets, he swamps you out anyway. So it comes down to (and this is why I think the talks this afternoon and this morning are related) when you use Pravin's numbers, you are not going to get anti-jam just from spread spectrum modulation itself, and you must try to put nulling and spread spectrum together. I still think it's a really tough problem, and I think that is a communications problem at the moment. We've talked about synchronization, we've talked about a number of things, I think it is a tough problem.

COMMENT-GOBIEN: I think there is a clear implication in all of this. Really, it is one of the results of Ed Speer's study, also, the only way that you are really going to solve this problem is to go out and kill a whole bunch of jammers first, and then kill the jam-resistant system. You have to assume that you cannot kill them all, and maybe you can afford the ECCM to resist what's left, but there is no way you're going to attack this problem with any kind of a realistic way, if you don't do something to get rid of a whole bunch of jammers, as you first said.

HUTH: In Pravin's satellite area, I can see how you might overcome that with the type of techniques that he talked about, if you make them all work. Tactical is really just a tough area.

BEDROSIAN: I'm interested, not so much from the strategic communication point of view, but from the tactical one. I want to ask the speakers in general how does all this look on the other side. What are their problems, and what are they doing about them? Or what ought we to do about them, or what should we be doing about them? Are we worrying about something that is unique to us? I won't mention who the other guys are.

HUTH: I know it's Kolmogorov, he told us this morning. (Laughter...)

McELIECE: I will just sneak in an academic-type remark. I've been asking that question in my own life for a while. By introducing the game theory, I've been trying to introduce a symmetry to the problem where the answer to one of the questions must also answer the other. So that this game-theoretic result (in the academic world as it may be) I mentioned a while ago answers the symmetrical problem for the jammer. "What's the worst thing that could happen to him?" And for the communicator vice versa. In open type conferences it's always struck me as at least asymmetric and possibly even strange that all of the remarks are about anti-jam, rather than anti-anti-jam.

HUTH: There's one other comment that goes along with what Seymour and Jerry said. The problem is that you are out there, and you are trying to communicate. The jammer is trying to stop you from communicating. The unfortunate part is that in this game he's also trying to communicate. He's playing his game. It is not clear when you have to play both of the games, that we are not trying to jam as well, even if it is friendly, or maybe if it

is even intentional. There's decoys and so on. It gets to be a much more complicated game than we've played here today In other words, both people are trying to do something.

QUESTION: A specific question. What frequency band do these hypothetical people use? Is it not UHF also?

ANSWER: No, it's VHF. So they might be safe there.

COMMENT-RISTENBATT: I'd just like to make a comment on that. I'd like to ask what the panelists think about this. Might it not be true that we have a higher reliance on communication systems? And we've built that into our systems, more than this Kolmogorov?

HUTH: Well, you were talking Kolmogorov, I think that your example was the Israeli war I don t necessarily think that there aren't some environments where it is. It depends on who's supplying the *equipment.*

COMMENT-GOBIEN: That's a hard question to answer. If you look at the doctrine, I think that's true. The Soviet doctrine is one of training pilots for instance (getting back to the tactical air problem) almost to be robots who execute instructions without exercising a great deal of judgment, without even being taught to exercise a great deal of judgment. So in that sense I suppose there is less of a reliance on free-wheeling communications with their colleagues who are also involved in this battle. On the other hand, they are critically dependent therefore upon some datalinks by which they get their instructions which they have to execute. So it's not really clear, that they are that much less reliant.

HUTH: I might make a comment that puts it in a different perspective. It's the assumption that there's the enemy over here, there are the good guys over here The enemy develops its communication systems, and the good guys develop their communication systems. In some cases (in recent wars we can talk about), the communications came from the same source. And in fact, we're in the environment where they were jamming, trying to communicate in the same band, and had been trained in the same way of carrying on a battle like Argentina-Britain. That's where I think with the game theory, if you really want to play that, maybe the approach has a larger dimension than you chose.

COMMENT-BEDROSIAN: I'd like to comment on that too A number of the studies that I've been involved in of Soviet Air defense in particular, suggest to me that they have a far more aggravated air defense problem than we do. They are preparing themselves against a large bomber penetrating force, cruise missiles, the sorts of things we're not really thinking of in numbers anywhere near theirs. I suspect that they have an air control problem that's more severe and requires better communications than we do. It might be quite different for other kinds of wars, if you are going to talk about a central war, than for something like the Israeli war, or Vietnam, or similar things, then that might not be the case. But I think they have important and serious problems. I just don't know what we're doing about that. This probably isn't the place to talk about it but I do think that viewing this thing symmetrically and letting that not escape our thinking is important because it influences how you structure your thoughts, and how you pursue different technical issues, much like you were saying.

LEVITT: Another thing, it also allows you to come up with the best ECM strategies. When you keep changing hats back and forth, when you're developing ECCM strategies and you play the game

theoretic aspects, it gives you benefits on both sides. The other thing that comes to mind, if you don't design for the worst case, if you don't allow the enemy, the adversary, whoever it is, to have at least as much smarts as you have, then you are really deluding yourself. When push comes to shove and you actually have to use this in a real situation, you can be dramatically let down, if your C-cubed support just disappears. So I think you really have to design from both an ECM and ECCM point of view and use that full-strategy approach.

HUTH: It's especially true when the people that are using the communications were trained in using your communications. This is the real thing where the stuff is deployed, and then it turns out who's on whose side again. It's a serious problem if you haven't done both sides of the fence.

GOBIEN: But there is a counter-argument, one which perennially bothers me. It has to do with the fact that we end up never fielding anything because we are always waiting for the better, and every time you get a third of the way there, there is a better solution on the horizon. So you drop what you're doing and pursue that. There was a beautiful example that I ran across recently that made me livid, but I really don't think this is the right environment for it. There are lots of evidences and indications around that if we fielded some real simple-minded ECCM systems, 10 years ago, the kind of stuff that nobody in this room would maintain is in any sense robust. (But at least it's something,) that we'd be miles and miles ahead of where we are.

HUTH: I think that's an ongoing complaint that's happened. It's a real research problem. That's part of the problem being the theorist versus trying to put the box out there.

PRICE: In radar there's an ECCM technique, I think it's called burn-through, which is just vast amounts of integration time until you see what you want to see. Somebody was talking about two-way networking and the use of adaptive transmitters possibly. It's futuristic, but could be very valuable. I wonder if there is anything that we have in the way of a communication burn-through capability now where the receiver, knowing that he's being jammed, just asks the transmitter to repeat over and over again until he gets the message, or he can get some information.

HUTH: There are strategies of that type. I don't know who wants to address it.

STEIN: Let me make a comment about adaptive data rate systems in general. Every time you see someone studying them, he has to assume, in order to make them work or in order to even analyze them, that the return link is free from problems. The fact is that the return link is never free of problems. If I remember in the past when people looked at adaptive data rates, for example, to work through fading, and I think jamming is a very similar situation, the return is not that great, and the complexity tremendous. There is also another problem. I don't know what the current military doctrine is now, but not too long ago, it was that if a communicator is being jammed he is not supposed to show by his reactions that he's being successfully jammed. He's supposed to keep doing what he's always been doing just to keep the enemy from knowing that he's been successful. That creates a lot of problems. The whole problem of how you convert from, call it a peace time solution to a war time solution, has plagued any kind of spread spectrum system because none of the military commanders really want these things around in peace time. They only cause

trouble.

HUTH: They may not even want them around then. What Jerry is saying is that people get very scared, and they don't want to communicate anyway if they think that they're going to put themselves in jeopardy at all. So they give it up. That's the first thing they think of. Trying to keep them communicating when you need to talk to them, is the problem.

STEIN: You really have to believe that if we are going to have these systems, they are going to have to be operating before the crisis actually occurs. Trying to coordinate afterwards is looking for real trouble.

PRICE: At least on the return link it is low data rate, because it is just binary, I've either heard of it or I haven't. The other thing is that maybe LPI techniques, or hiding a signal under your own overt signal, saying that I am being jammed, wouldn't give it away, if you are being jammed.

HUTH: The only thing is that the forward link you selected, as you watch the numbers go by, is very low data rate to begin with. The return link is of the same order.

PRICE: I thought the forward link might be voice which is high data rate to me.

HUTH: But not always.

LEINER: There is technology coming down the line that will support adaptive data rates, adaptive coding, a lot of these things, but as you'll see, we'll be addressing it somewhat tomorrow in the spread spectrum network session. The problem is how do you control it. It's not so much how to establish the communication between the individual nodes, but it is what protocols, what algorithms, do you build on top of that so you can do something intelligently. For example, in packet radio we've had power control, the capability to do power control since day 1. We still, here we are 10 years later, we don't know how to use that.

WOZENCRAFT: Just to comment along those same lines, if you have any kind of a symmetrical link, about the same kind of data rate availability in both directions, both directions being jammable, but also both directions having a capability for error correction-detection coding, then you can come up with two-way strategies, which at least in theory are pretty much fail proof.

STEIN: I did not mean to comment that you cannot come up with strategies. It is just that when people have worked out the details, my impression has been that generally the gain, the effective information throughput gain, was not that high when you finished allowing for all of the problems that might occur. It's that kind of difficulty.

WOZENCRAFT: I'm not sure I've encountered all the problems that could occur. But it seems either typically you can think of things like 40% throughput in reasonable situations. Which isn't too bad in a coding world.

HUTH: Thank all the speakers, you did a good job.

SESSION 3 - APPLICATIONS OF CODING TO SPREAD SPECTRUM COMMUNICATIONS

LLOYD WELCH: The title of this session is Applications of Coding to Spread Spectrum Communications. Our speakers are going to be Elwyn Berlekamp. Joseph Odenwalder and Jim Omura. To me the title Application of Coding to Spread Spectrum is a little bit broader than what my speakers are going to be talking about. I feel that there are really two aspects. There's the error correcting coding, and there's also coding to spread the spectrum. All of my speakers have chosen to speak on the first subject, the error-correcting coding. I don't know, perhaps they feel there are no problems in the other area, or its trivial. Maybe you feel that way too! I noticed yesterday, there were several speakers who in running through the options mentioned direct sequence and frequency hopping and that was it. There were no alternatives. Well I pose that question. Are there other methods of using the sequences to spread a spectrum? Also, maybe you feel that there's no further problems involved in the sequence design for frequency hopping or direct sequence. Maybe that's true in the jamming situation, but I feel that there are other situations, for instance in spread spectrum for multiple-access in which there are still problems in designing sets of sequences, especially in the frequency hopping situation. So I feel that there are some problems in this area, but nobody is going to talk about them today. (Laughter...)

Let me return to error-correcting coding and introduce our first speaker, Elwyn Berlekamp, who received his Ph.D. at MIT in 64, and then spent 5 years at Bell Labs. He's been a professor at Berkeley for the last 12 years in mathematics, EE and CS, and of course he's now president of a thriving company, Cyclotomics. He is also author of a book, Algebraic Coding Theory which is certainly relevant to today's talk, and he's a co-author of Winning Ways which is also relevant to spread spectrum communications in the jamming environment.

ELWYN BERLEKAMP:

I think what I will talk about has obvious applications to pulse jamming although that isn't the way I got in to the problem, so maybe I should tell you the motivation which got us started. There is a traditional sort of folk-theorem wisdom which is that if the channel is bursty, then you should whiten it, and the way to whiten it is to figure out what the longest burst lengths that you're likely to see is, and then interleave by either direct interleaving, or block interleaving or pseudo-random interleaving, RAM interleaving, or whatever, to some depth far bigger than the maximum burst length on the channel. Then the equivalent channel is white like the case you've analyzed and you charge on merrily away

I've found a number of applications where this is sort of procedure is not possible because originally the cases of most interest to us came about where there was an external system constraint on delay. For example, there's going to be a voice communication link and any delay more than a couple of tenths of a second is intolerable. On the other hand, the channel's burst distribution was such that, sometimes you get bursts lasting longer than a couple of tenths of a second Apparently, your communication link will fail when this happens, but you also will be getting lots of other bursts of shorter lengths, e.g., sometimes a tenth of a second, sometimes bursts of length 15 milliseconds, and a wide distribution of

burst varieties. So the issue arises: The given delay spec does not allow interleaving to a depth necessary to carry out the whitening process. Originally in these cases, the delay spec turned out to be much more of a driving force than the memory requirement for interleaving, and anyway, we all know that memory is getting cheaper and cheaper, and that's usually not where the issue is any more.

Suppose you have a delay requirement and yet you are going to have bursts fairly long with respect to that. How do various types of interleavers perform when you hit them with bursts longer than they are really able to cope with? So Po Tong and I analyzed every system we could think of and ended up inventing a new one that we believe is significantly better. Let me show you the bottom line. See Figure 7.8 (This is the end of the talk rather than the beginning, but if I show it first, I'll get everybody's attention, I think, and the rest of the talk will consist of backing off slightly from all these claims. Laughter...)

Figure 7.8 is a plot performance for a variety of combinations of coding and interleaving schemes. We're dealing in this case with 5 bit characters transmitted sequentially so that every 5 bits: i, i+1, i+2, i+3, i+4, all go together as they wander across the channel and through the interleaver. The primary code that was intended originally is the (32,24) Reed-Solomon code. It's actually extended slightly but it's essentially a Reed-Solomon code, similar to the JTIDS RS code except that the rate is higher. The rate is now 3/4ths and you use this code with pseudo-random interleaving and a delay constraint which is the same for all code words. I must confess, I have forgotten the precise value of the delay constraint; it is approximately two thousand bits. About 5000 bits of delay constraint would normalize out the speeds.

The question is the following: If the system, whatever system it is, gets hit with a burst of length n characters, where n is 100, 200, (300 characters, 300 characters is 1500 bits, so needless to say, that's going to bring us down) then, what is the expected number of additional decoded bit errors that get through the system?

Originally we did this in the case when the background error rate was adjusted so that in the absence of the burst, the output bit error rate was 10^{-6}. It turned out that there's not much difference if you assume that there are no errors in the background at all. A random error rate of 10^{-5} is low enough that the channel might as well be error free. That's a much easier case to analyze so maybe we should think about the background as error free. Now along comes this burst of n characters long, and as a function of n, within the burst the bit errors are totally random (that's the Gilbert type of noise model). You compute an expectation taken over all phases in which the burst might begin relative to your interleaver, and you average over the detailed noise pattern in the burst and then you get these various curves of Figure 7.8.

System 1 is the new one, that is, the one we are pushing today. Of course, I'll talk more about it later. It's apparently much better than anything else. Even with a 200 character burst, we expect to see only 100 bit errors out. These other schemes have 400 or 500 output bit errors in the same situation, or near to 4 times as many errors as you're getting out of (1) at that burst length.

The curves (5) and (6) in Figure 7.8, labeled CIRC, are an actual system that's now in the field, implemented in the compact disc business. I don't know if you're following Consumer Electronics, but Sony and Philips have introduced a product called the compact disc. It's

digital; it's written with lasers on an optical media which is read in your player. You can buy a player for $800 in Japan, and you can buy the entire library of compact discs for about $10 each. The last time I checked, there were only about a dozen such discs available, so you can buy all the titles. That's really what's holding up this venture: The stamping process is so bad that they didn't have adequate yield on the number of titles they want (and they were eager to make a lot of money last Christmas). But inside this device is a 1 IC gadget, actually its 2 ICs, one of the ICs is a 16K RAM which is a conventional off-the-shelf device. The other one is a custom chip that Sony and Philips have which they call CIRC (The CI stands for Cross Interleave. R stands for Reed-Solomon, C stands for code). They are very proud of their CIRC, which in fact, has these parameters, but gets there in what I regard as an unfortunately complicated and confusing way which chews up much of the area on their chip, and cuts the speed of their custom chip down to 2 megabits. This is fast enough for their applications but with these parameters you could go much faster by doing the whole thing differently. They are also proud of their interleaving which is, as I showed here, the worst of all in this comparison. The code is capable of operating on curve (5) which isn't quite so bad, and they have published some papers mostly by Vries at Philips who has discussed this. The system uses two RS codes and concatenates them, the first one being (32,28) and the second (28,24). By concatenating them in parallel, they end up with the performance shown. It is disappointing but it's on one chip which is inside a product now in production, so it attracts a lot of attention. It's been pushed heavily by Sony and Philips in a lot of other applications but I'm not a great fan of it as you can see.

Anyway, let me tell you about helical interleaving since that's really (in my view, in this application) the way to go. I'll also mention some caveats. Since I think helical interleaving is significantly better than pseudo-random interleaving, I think it raises a real issue in the jamming environment: Do you want to assume the enemy is pulse jamming or not, and if he is, then you want to go off in the same kind of direction that Dr. Levitt was discussing yesterday (session 2)? If you know what the enemy is doing and you can count on him doing that, then you can do much better than if he is min-maxing you somehow. In terms of min-maxing on interleaving, I'm not aware of any serious studies on this. It's sort of trivial on all the applications I've seen for the jammer to build his own interleaver which matches yours and put his jamming on the upside of that. So he wipes you out no matter what you are doing, assuming he knows what you're doing. And everybody says well I'm doing this pseudo-randomly On the other hand the pseudo-random stuff is wired in and not classified and I don't see it as all that hard for him to copy that! The only way you are going to defend effectively against the smart jammer is by putting crypto-variables into the interleaving in a rather intricate way This isn't done as often as it might be. Now for the pulse jammer, you can pick up quite a bit of performance with helical interleaving. My original motivation was not against pulse jamming, but against natural noise sources which in many cases are just bursty by their nature and so that whole philosophical issue doesn't arise We just concentrate on correcting the bursts.

Most of the interleavers I'm talking about can be implemented as shown in Figure 3.1, with 1 RAM. This differs quite a bit from Ramsey's paper which is the most recent serious study on interleaving of which I'm aware. That came out in 1970 and there he studied interleaving where

the game was to minimize the memory. He ended up with the assumption that all your memory should be delay lines. From a brute force technology point of view, that's probably right. There's less silicon area in the delay line than in the RAM. However, the market for RAMs is so huge and so many people are building them, that the price of RAM is much less than the price of delay lines. In fact the cheapest way to implement most delay lines is with RAMs. So we investigated 1 RAM interleavers. There's the RAM, data comes in, data goes out and all the guts are in the address sequence, which jumps around according to some pattern or other. The most effective ones which squeeze the memory down the most, all have the property that on a single cycle, you do both a read-out and a write-in. So when an address comes up, the old data gets taken out of there and the new data gets put in, and then the address jumps somewhere else. That means the RAM is always full. Now this puts constraints on what types of interleavers are doable, although most of the ones I have talked about are in fact doable in this format if you are clever enough about how you pick the address sequencing.

In general terms, co-ordinatizing is defining how the RAM interleaver system works. There is a geometric address pattern that you can follow or define and you can also define it with algebraic equations, but the exposition of such detail lies beyond the scope of this overview.

The interleaving depth in all of our "pure" helical interleavers is one less than the block length, so the codelength 32 would have interleaving depths 31. That can be adjusted by cascading interleavers, in this version. If you want to have a long code, say code length 255 interleaved to depth 16 or something as opposed to depth 254, that's doable by cascading interleavers. You use a conventional block interleaver with much smaller parameters and then a big helical interleaver which works on characters which are much bigger. The characters going into the helical interleaver are larger blocks that came out of the block interleaver. This gets around the problem of having an interleaving depth constraint less than one block length. The price however, is that the block interleaver does not have anywhere near the performance of a helical interleaver. Roughly, the block interleaver gives you a depth either a half or a quarter of what the helical interleaver would for the same delay. There's a deep and a shallow version of these staggered interleavers, depending on the relation between block length and interleaving depth. If you have a code length of 255, and you want an interleaving depth of say 16, then that's what I call deep see Figure 4.1. If you want a code length of 255, and interleaving depth of 10,000, then that's what I call shallow see Figure 4.2. The difference is just to which side of the helical interleaver you connect the block interleaver. In both cases you can arrange this so that the helical interleaver has the bigger delay and the bigger memory.

The factor of 2 or 4, (I'll discuss the difference later), that you lose because a block interleaver (like pseudo-random, or even all these other non-helical schemes) is inherently weaker, depends on your assumptions about fade detectors. Let me tell you why it is that helical interleaver performance is so good: There are two reasons. One is that we're using a scheme called burst forecasting which works only, as far as I know, with helical interleaving. And that picks up the last factor of 2. If you don't use burst forecasting, you might use an ideal fade detector, or an ideal jam detector, or whatever it is. This would be a gadget stuck on the modem, that converted all your error bursts into erasure bursts and told all the codes, "Gee, we're in the erasure mode now". If

you put the ideal fade detector in all the coding systems under comparison, then the helical interleaver is still the best, but the difference is less massive.

The performance of several interleavers with an ideal fade detector is shown in Figure 6.6. With no interleaver, you get some phenomena at very low burst lengths because of the code itself. All of this is for a Reed-Solomon (32,24) code, but similar qualitative shapes occur with other codes. (We've done 4 or 5 examples. They all seem to look the same). You get some peculiarities at short lengths because, until you are up to the burst length that's sufficient to override the code redundancy, you basically have no errors. We're showing a little error there due to the background (we are assuming the background rate is sufficient to cause a 10^{-6} output rate without any erasure burst). Along comes the erasure burst, that starts degrading things not because the erasure burst is enough to cause troubles, but because the erasure burst is such that the background error rate starts getting through. Then when the burst length exceeds the code's capability, the number of bit errors jumps up and continues in a straight line (with no interleaving).

If you use a block interleaver of depth 16, you get the kind of funny piece-wise linear phenomena shown in Figure 6.6. These lines and break points have to do with when you reach the point that you're starting to get enough errors in one code to push it over its error-correcting limits. Figure 5.3 is the curve used in the construction of these interleaver performance comparisons. This is the performance of the (32,24) Reed-Solomon code with a fade detector, s erasures and an input bit-error rate in the background. It's easy to see which piece-wise continuous curves occur. If you have 5 erasures plus some background bit error rate then you're on the s=5 curve but when the erasure burst gets long enough to start hitting curves with s=7 instead of s=6, then you jump up to the different curve indicating poorer performance.

Now pseudo-random interleaving is not so bad really, as shown in Figure 6.6. It turns out that pseudo-random interleaving is better than block. But, helical is better yet in the large range of most interest. The problem is that once you get up to the point where you're just barely overtaxing the system, then in fact helical is worse than pseudo-random because every helical block is getting zapped just enough to knock it out. The pseudo-random one on the other hand doesn't see that much difference in small variations about this critical burst length because in either case it is getting some and missing some. Both of these curves assume an ideal fade detector.

If you take the ideal fade detector away then you get a different curve see Figure 7.6 and basically everything shifts over; it's much worse now with no fade detector. Basically the number of characters in a burst which can cause you trouble, got divided in half. But the beauty of helical interleaving is that you can get performance almost the same as if you have a fade detector even though you don't have a fade detector. And the reason is that unlike block interleaving, pseudo-random interleaving, and any other schemes, you can forecast the bursts. [This technique was not used in Figure 7.6.] and this is a fairly easy thing to do in software. There are a number of ways you can tweak your algorithms to do it quickly. But the crux of the matter is this, it's in an intrinsic advantage of convolutional codes in general that they can start by assuming that if you assume the past is error-free, then you've already got a big head start on decoding the next bit because you've already decoded the past.

The same sort of thing is true in helical interleaving. If I assume that I've already decoded the past and noted that a burst was present in certain positions of one decoded interleaved word, then jolly well, it's very likely to be continuing on, so it's probably going to continue on into an adjacent interleaved word which is yet to be decoded. So we can forecast the burst and erase those corresponding positions in the next word. And so this is what I call burst forecasting.

There are some subtleties. It's the familiar weather-prediction scheme. The best weather prediction scheme until the last few years has been: you predict tomorrow's weather to be the same as today's, and it works in almost all climates, in almost all seasons. It's a great system. (Laughter.) And that's really all we are doing. If you're in a burst mode, you project you're going to stay there. If you're not in a burst, you project, "There ain't going to be one." That means when the burst comes along it hits you by surprise, but only in the beginning, and when the burst finally goes away, you've hurt yourself by forecasting it's going to be there when it's not. But if the bursts are long, these edge effects don't count for much and details of how you handle these effects don't matter too much either.

There are a variety of forecasting schemes. The one we are doing now enters the burst mode only when we get two character errors in a row. If we have an error twice in a row, then we say it's going to continue in that mode. If a single error-free symbol comes through in the burst mode, we think maybe that is an accident, we don't believe that yet. That is, if you handle burst ... burst, burst, burst, error-free, you say, "Oh well, it's probably going to be a burst". But if it's error-free twice in a row, then you're going to say that burst is going to go away. And then it comes back. And yes, if you want to be cynical, there is a pessimistic symbol-error probability level for the bursts, where all the bursts have about 5% error rate within them, which causes the most trouble. That keeps us in an uncertain situation, in which we're always guessing whether or not the channel is going to change state. But if somebody comes in and tells us the symbol error probability in a burst, then we can tweak our algorithm and get ahead of them again. Anyway, that's the crux of burst forecasting with helical interleaving.

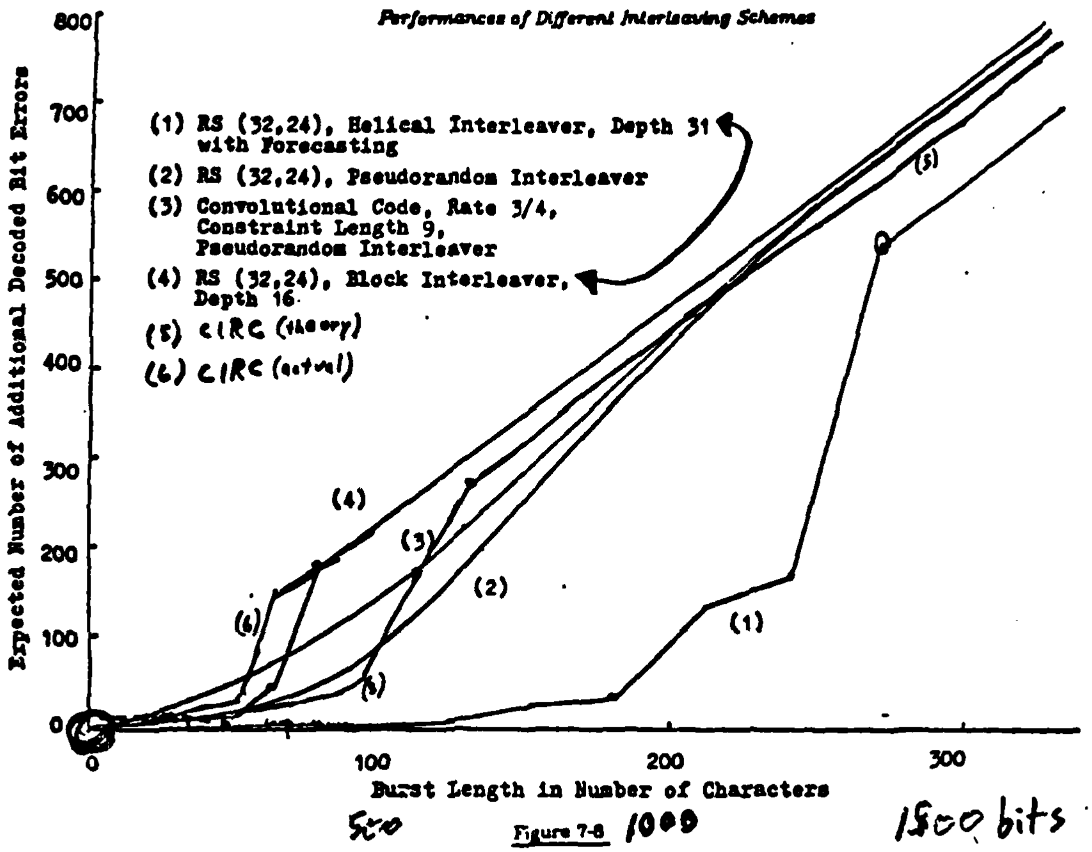


Figure 7-6

INTERLEAVING

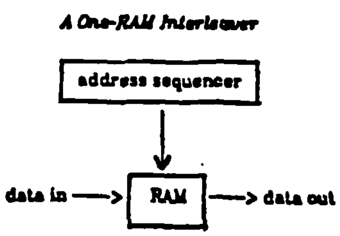


Figure 3-1

Deep-Staggered Interleaver

TRANSMITTER:

... —> Encoder —> Block —> Helical —> Modems &
Interleaver Interleaver Channel

RECEIVER:

Channel —> Helical —> Block —> Decoder —> ...
& Modems Interleaver Interleaver

Figure 4-1

Shallow-Staggered Interleaver

TRANSMITTER:

... —> Encoder —> Helical —> Block —> Modems &
Interleaver Interleaver Channel

RECEIVER:

Channel —> Block —> Helical —> Decoder —> ...
& Modems Interleaver Interleaver

Figure 4-2

Performances of Interleavers with Ideal Fade Detector

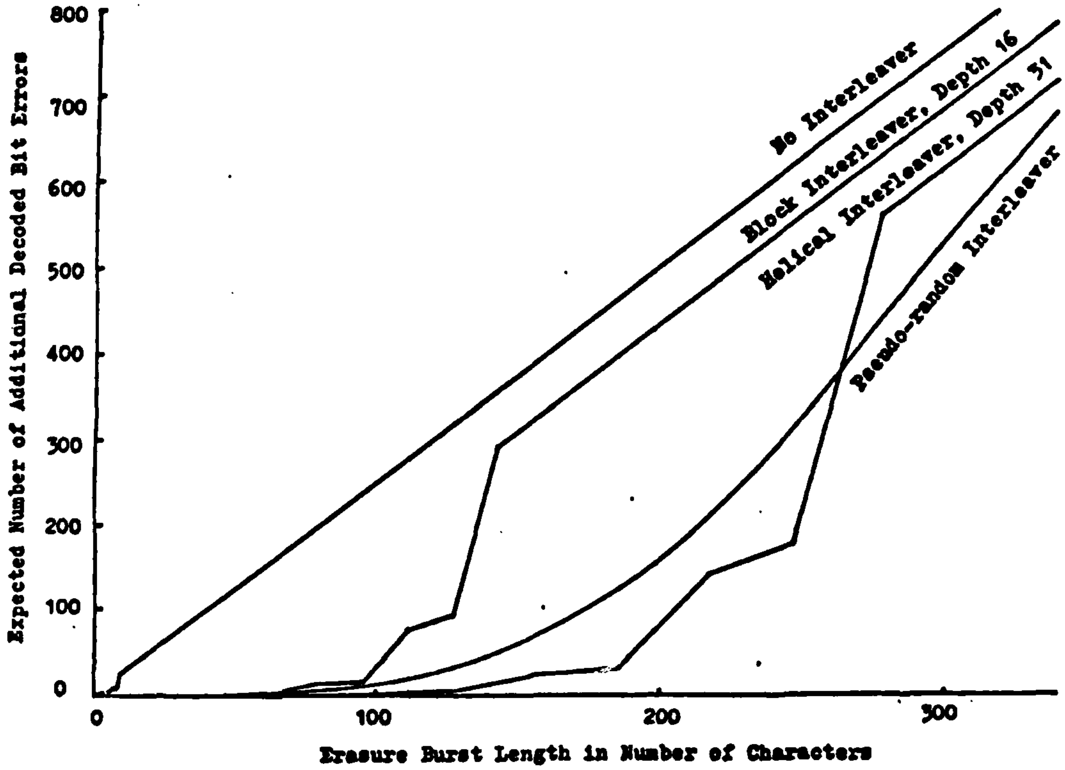


Figure 6-6

Performance of RS(32,24) with ideal fade detector and s erasures

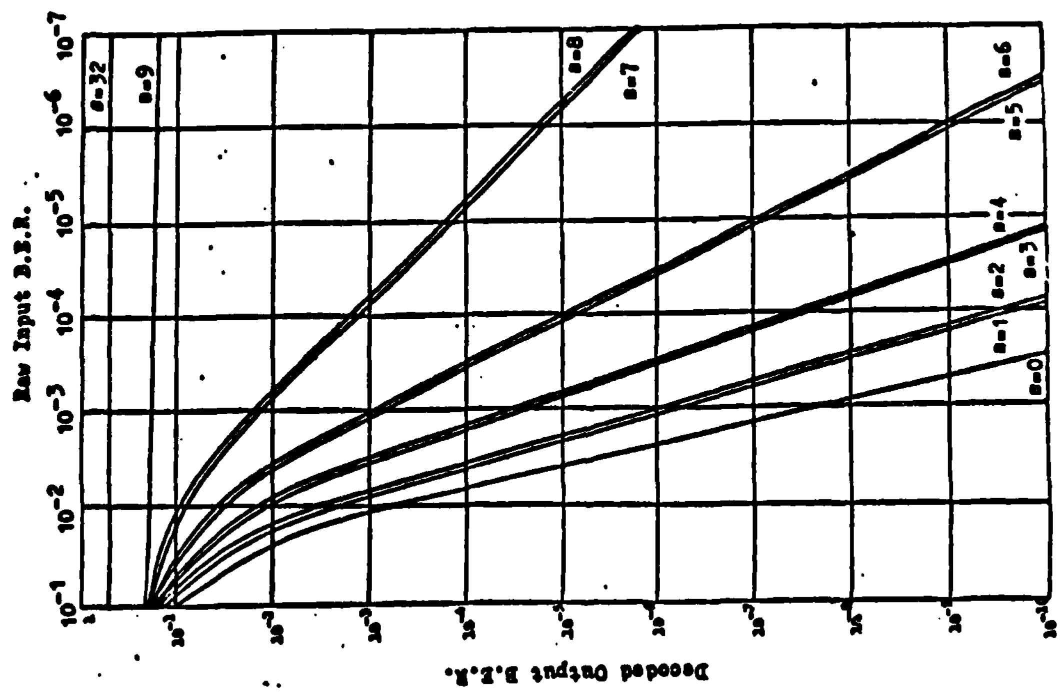


Figure 5-3

Performances of Interleavers with No Fade Detector

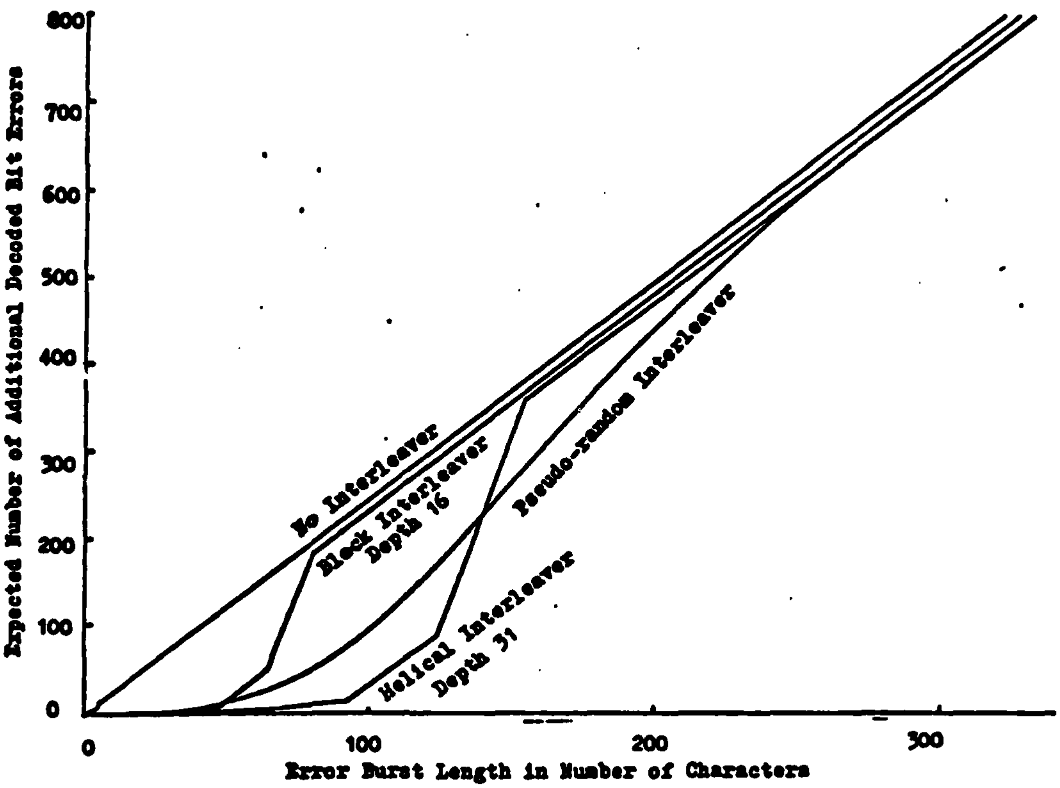
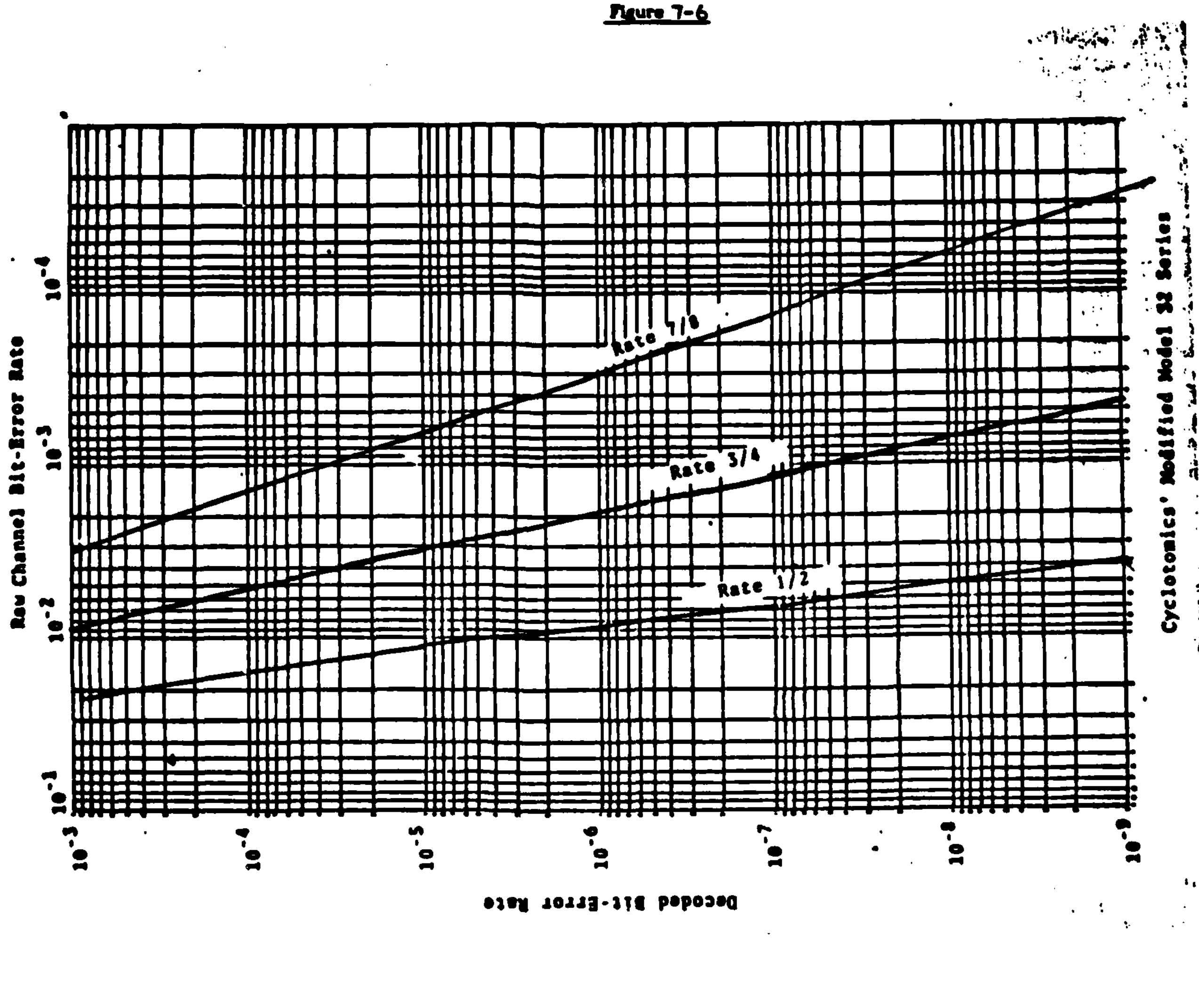
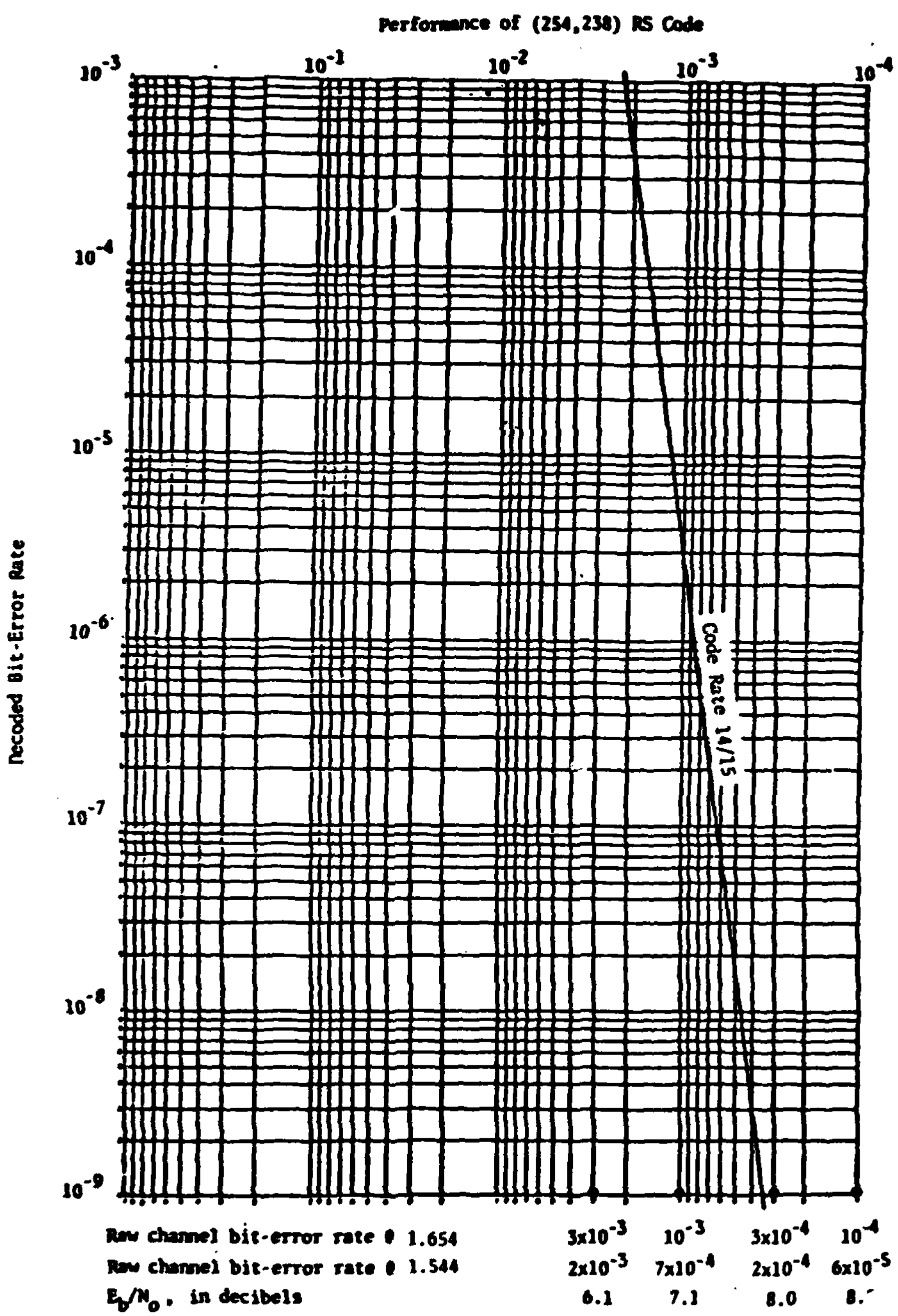
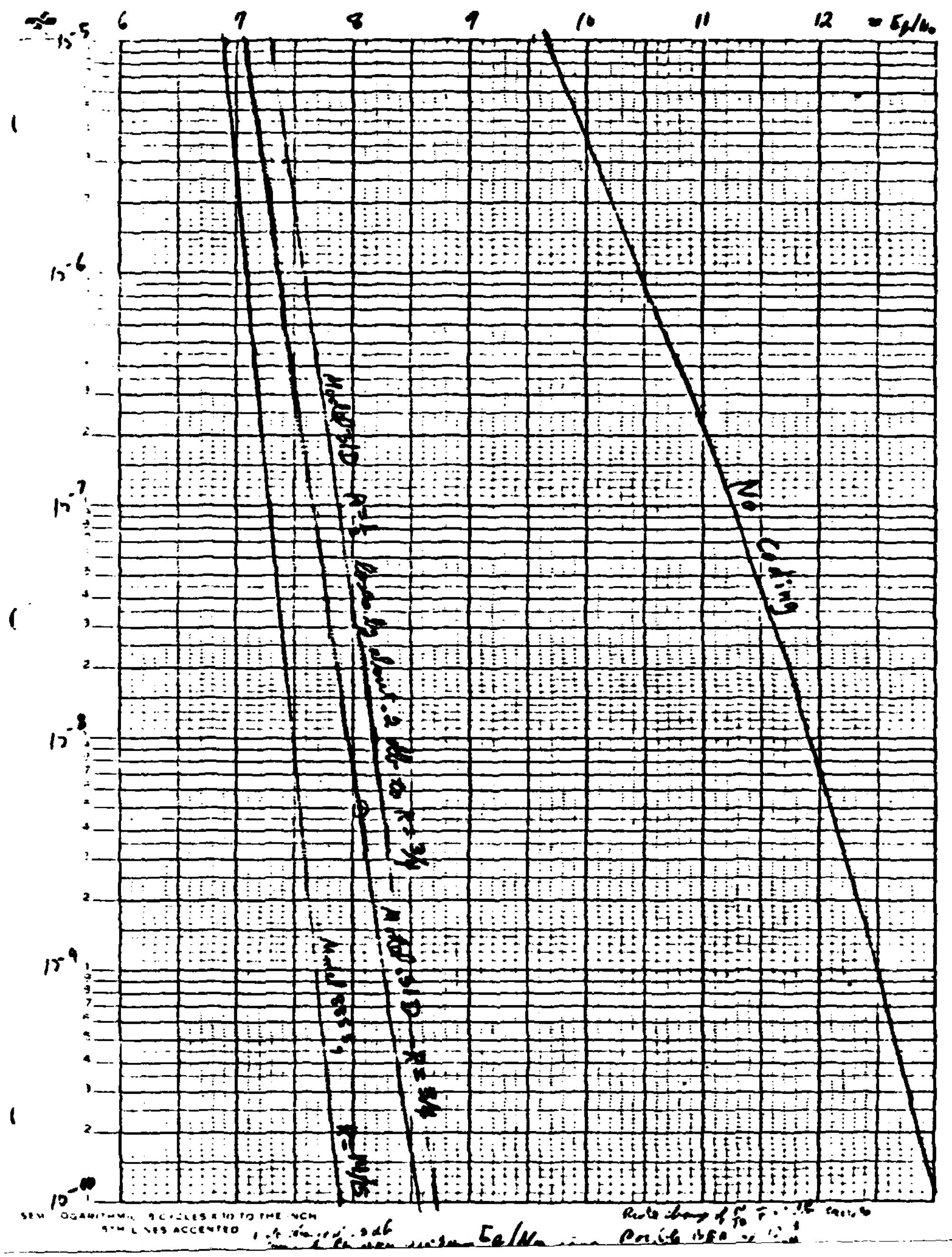
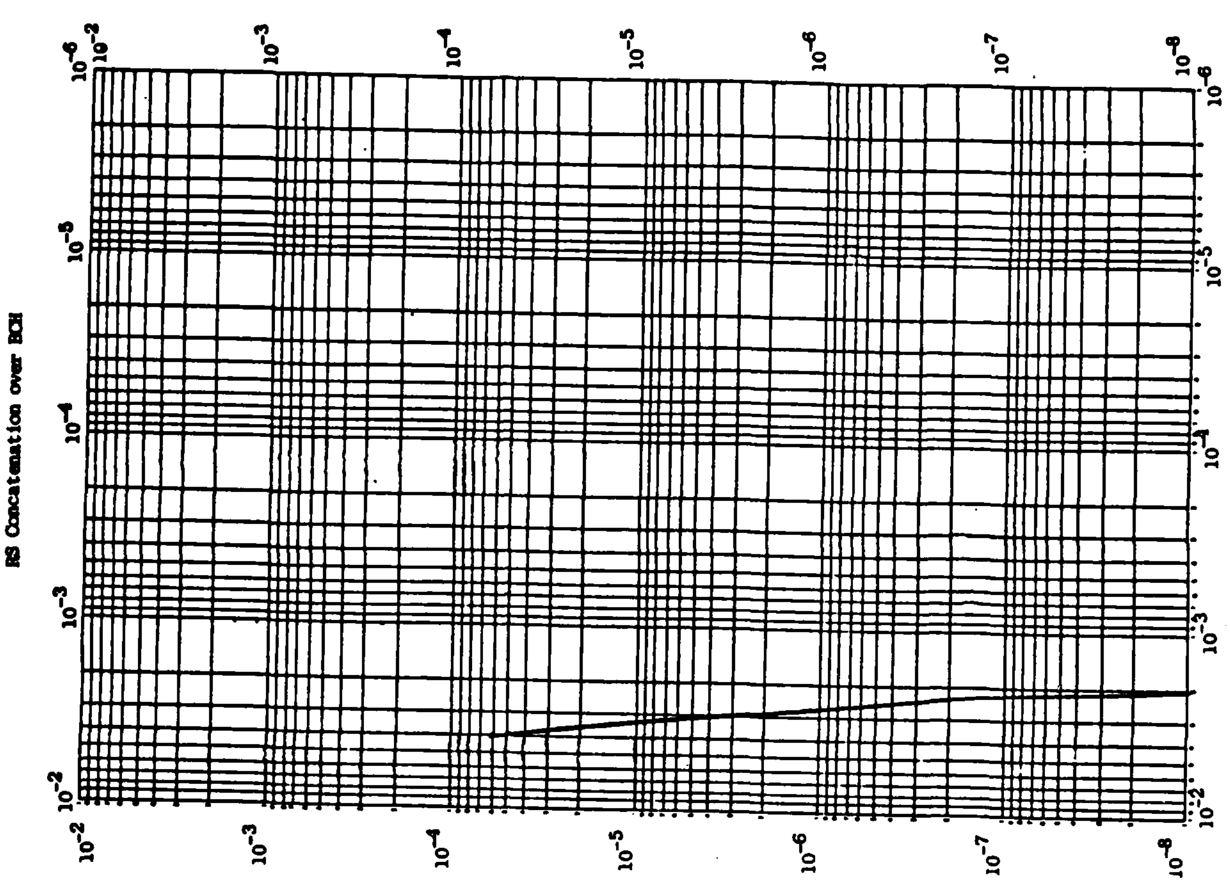


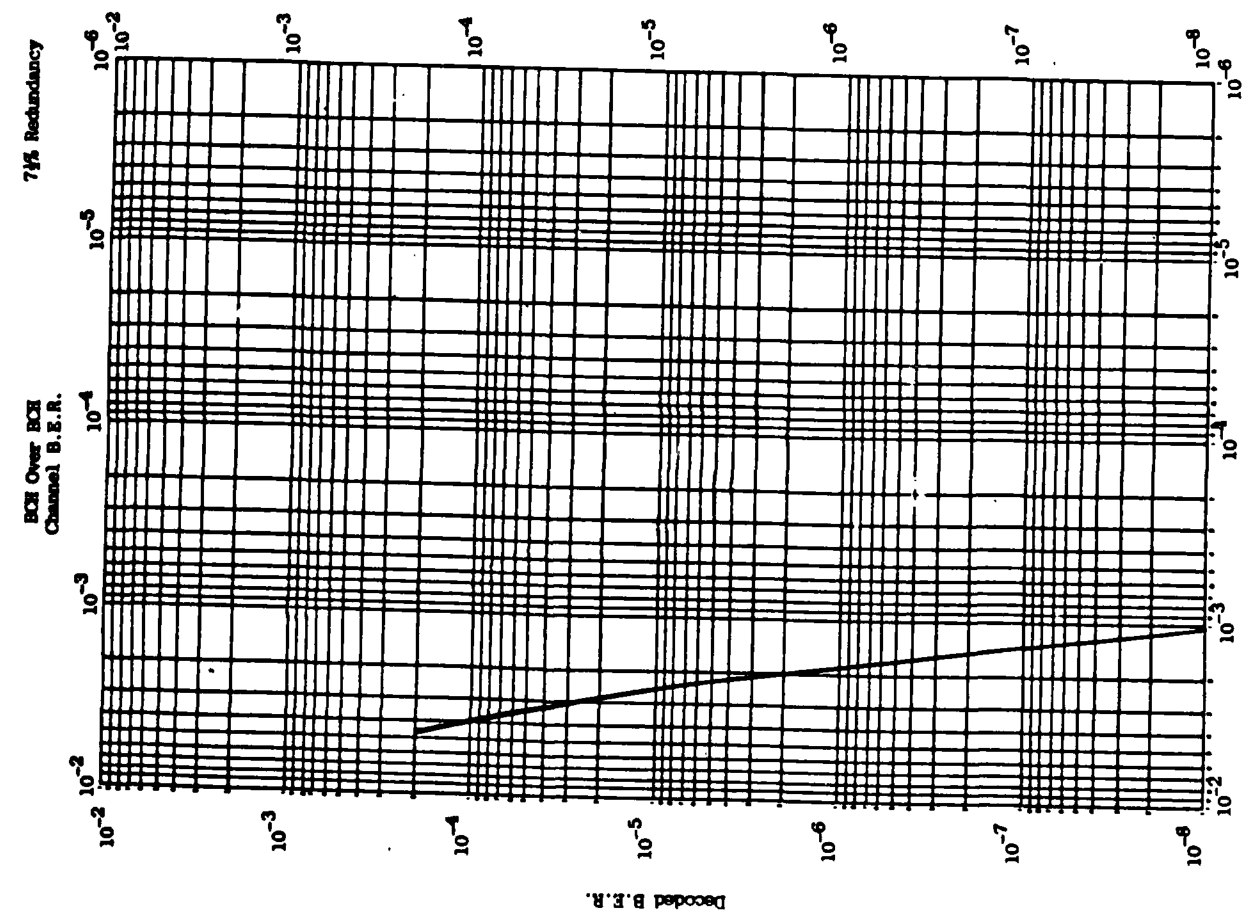
Figure 7-6

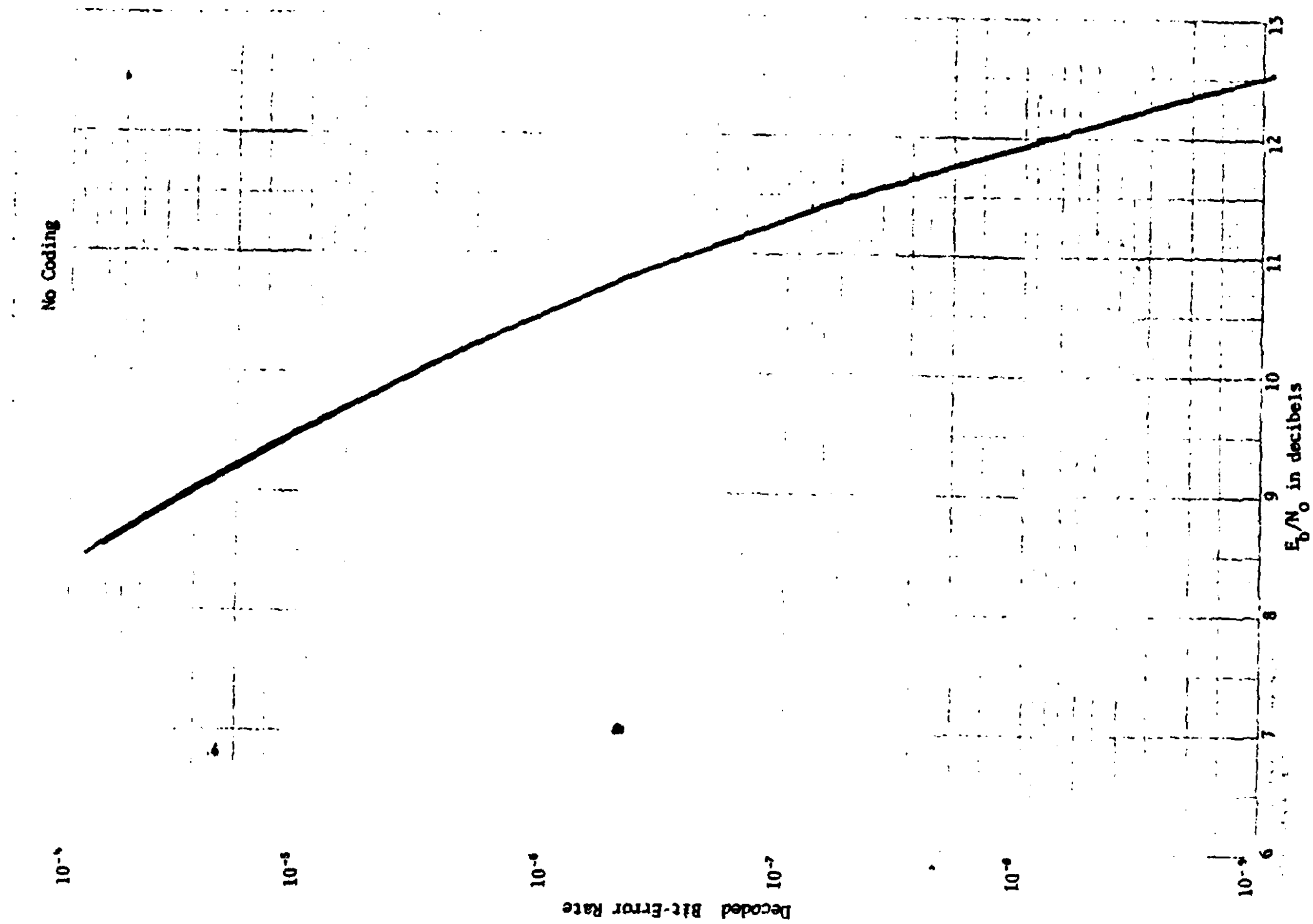


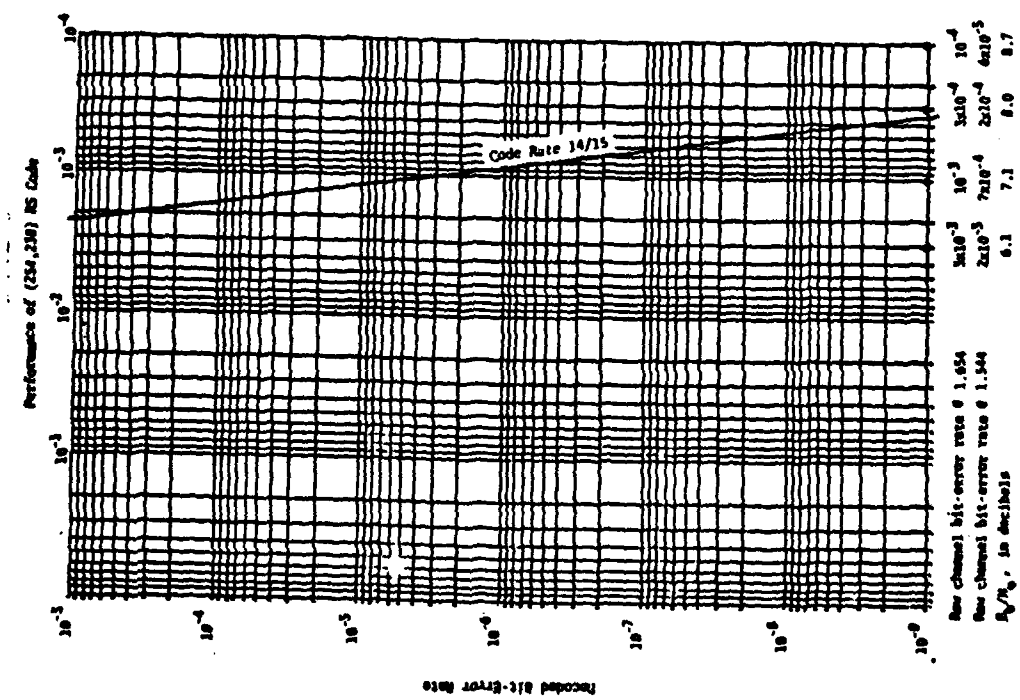

Performance of (254,238) RS Code
Decoded Bit-Error Rate
Code Rate 14/15
Raw channel bit-error rate @ 1.654
Raw channel bit-error rate @ 1.544
E_b/N_o, in decibels

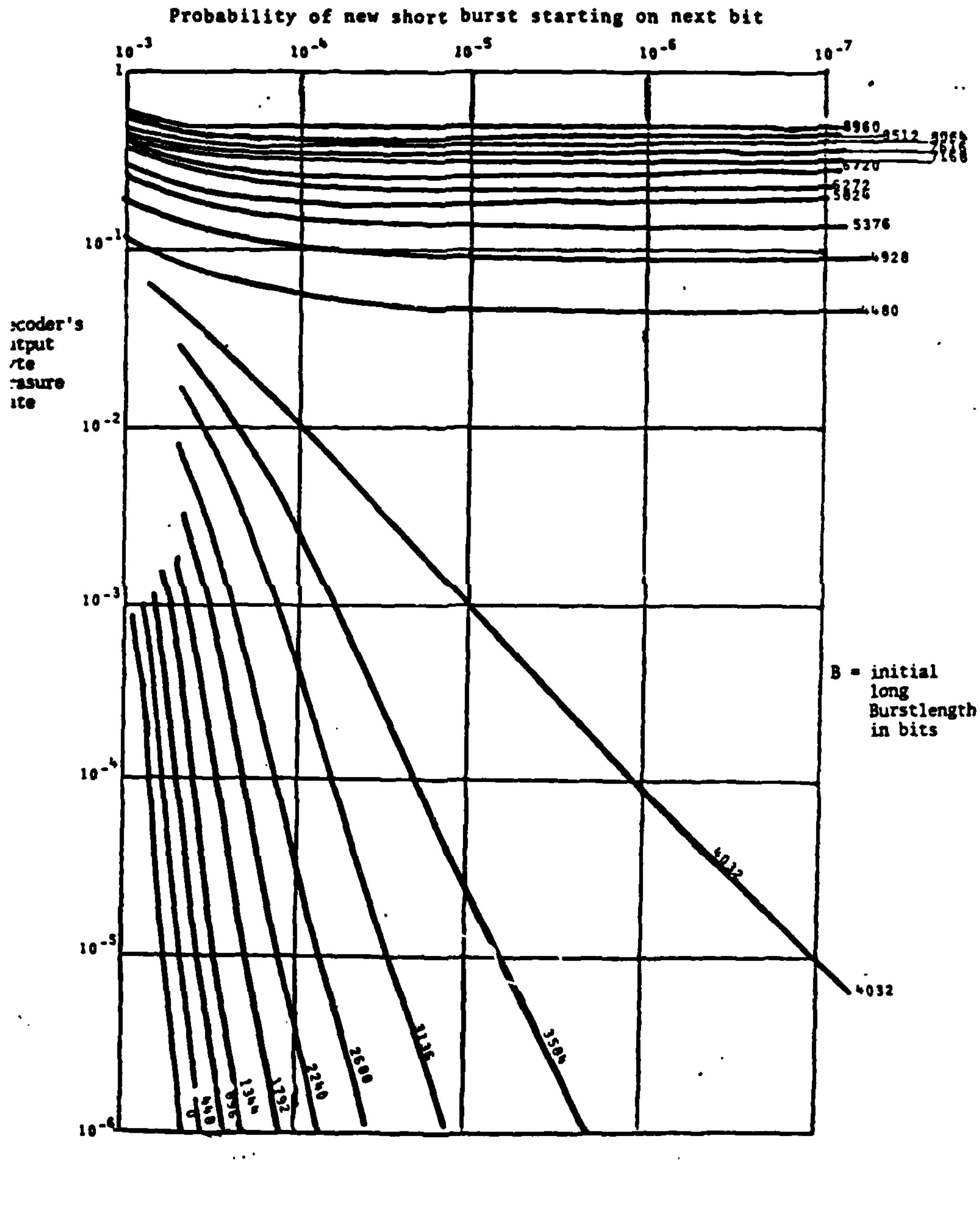

No Coding
Model 31D R=1/2
Model 31D R=3/4
R=14/15
E_b/N_o


RS Concatenation over BCH


7½% Redundancy
BCH Over BCH
Channel B.E.R.
Decoded B.E.R.


No Coding
Decoded Bit-Error Rate
E_b/N_o in decibels
10⁻⁴
10⁻⁵
10⁻⁶
10⁻⁷
10⁻⁸
10⁻⁹
6
7
8
9
10
11
12
13


Performance of (254,238) RS Code
Code Rate 14/15
Decoded Bit-Error Rate
Raw channel bit-error rate @ 1.654
Raw channel bit-error rate @ 1.344
E_b/N_o, in decibels


Probability of new short burst starting on next bit
10^-3
10^-4
10^-5
10^-6
10^-7
1
10^-1
10^-2
10^-3
10^-4
10^-5
10^-6
ecoder's
itput
yte
rasure
ite
8960
8512
6720
5376
4928
4480
4032
4032
3584
2688
2240
1792
1344
896
448
0
B = initial
long
Burstlength
in bits

8. REFERENCES

Berlekamp, E. R., 1963 A Class of Convolution Codes, *Inform. Control*, 6:1-13.

Berlekamp, E. R., 1968 "Algebraic Coding Theory," McGraw-Hill.

Berlekamp, E. R., and Ramsey, J. L., 1978 Readable Erasures Improve the Performance of Reed-Solomon Codes, *IEEE Trans. Inform. Theory*, IT-24:632-633.

Berlekamp, E. R., 1980 The Technology of Error-Correcting Codes, *Proc. IEEE*, 68:564-593.

Brayer, K., 1975 "Data Communication via Fading Channels," IEEE Press.

Carlitz, L., Roselle, D. P. and Scoville, R.A., 1971 Some Remarks on Ballot-type Sequences of Positive Integers, *J. Combinatorial Theory, Ser A*, 11:258-271.

Clark, G. C., Jr., and Cain, J. B., 1981 "Error-Correction Coding for Digital Communications," Plenum Press.

Elias, P., 1955 Coding for Noisy Channels, *IRE Conv. Record*, pt. 4, 37-46.

Feller, W., 1967 "An Introduction to Probability Theory and Its Applications," John Wiley & Sons.

Forney, G.D., 1971 Burst-Correcting Codes for the Classic Bursty Channel, *IEEE Trans. Commun. Technol.*, COM-19:772-781.

Fray, R.D. and Roselle, D.P., 1973 On Weighted Lattice Paths, *J. Combinatorial Theory, Ser A*, 14:21-29.

Lindsey, W.C., 1972 "Synchronization Systems in Communication and Control," Prentice-Hall.

Ramsey, J. L., 1970 Realization of Optimum Interleavers, *IEEE Trans. Inform. Theory*, IT-16:338-345.

Ziv, J., 1962 Coding and Decoding for Time-discrete Amplitude Continuous Memoryless Channels, *IEEE Trans. Inform. Theory*, IT-8:S199-S205.

Ziv, J., 1966 Further Results on the Asymptotic Complexity of an Iterative Coding Scheme, *IEEE Trans. Inform. Theory*, IT-12:168-171.

Ziv, J., 1967 Asymptotic Performance and Complexity of a Coding Scheme for Memoryless Channels, *IEEE Trans. Inform. Theory*, IT-13:356-359.

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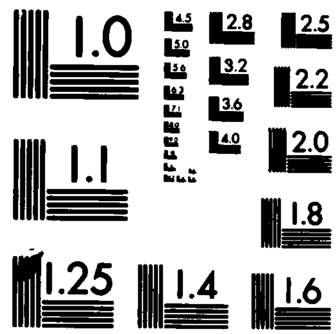

MICROCOPY RESOLUTION TEST CHART
NATIONAL BUREAU OF STANDARDS-1963-A

WELCH:

Our next speaker, Joe Odenwalder, received his Ph.D. at UCLA. His research interests include convolutional coding and concatenated Reed-Solomon and Viterbi decoding. He worked at Hughes Aircraft until 1972, and since 1972 he has also been with a thriving company, Linkabit, where he is an Engineering Director.

JOE ODENWALDER

I'd like to talk about some of the needs and long term trends in convolutional coding, and then quickly get into a few selected areas where I feel that more work is required. First I want to look at some of the system needs so we can concentrate on the things that are really needed see Figures 1 and 2.

Typically the first thing that comes to mind of course is performance. The main thing that I wanted to point out here, that has already been noted in several of the other talks, is that it's not just additive white Gaussian noise that must be considered. You have to make sure the system works in all the expected channel environments which may also include jamming, atmospheric simulation and, more recently, nuclear simulation. In some cases you might also have to contend with multipath and if you're talking about a satellite system, the uplink and downlink environments must be considered if they're both limiting performance. Right after performance the project engineer or the system engineer is going to be concerned with size, weight and power, and of course, cost. I'll just go over some of these things quickly and then discuss what we can do to improve things.

Continuing on with the needs, we need something that's very flexible. It has to be good for various modulations in some cases and I've noted that probably the most common types of modulation in the future will be where we'll have more and more frequency hopping systems as opposed to PN, again primarily because of the fading and nuclear simulation problem of tracking in that type of an environment. Probably the most common type of modulation will be MFSK for the lower data rates and for the higher data rates DPSK, again with frequency hopping. It would be nice to have some kind of coding system that's good for both of these types of modulation. Also, as I've already noted, it should be good for all types of environments. You shouldn't come up with some clever scheme that is good against tone jamming, for example, and have it be such that the jammer can take advantage of some particular characteristic that you've used to really shut down your system. Some of the other things you have to worry about are AGC or any kind of other information that you are using to do something clever to mitigate against some of the jamming environments.

Another thing you have to worry about is how to handle data from various sources. I'm thinking now of a downlink where you're receiving data from many separately encoded sources and you don't want to have a separate decoder for each separately encoded source. You'd like to have one decoder that you share and with convolutional encoding you have the path histories and metrics that you have to worry about. You need a good way of handling this. Even though this is an AJ system, people always want higher data rates. We've already mentioned synchronization, and in some cases node sync to worry about for convolutional encoding system, or even phase ambiguity resolution in some cases. I've also included interleaving in the coding part of things. The main interleaving restriction is usually not due to the size of the memory but to the throughput delay and synchronization time.

Where do I think things are going to try and achieve some of these system needs see Figures 3 and 4? Well, for one thing there's a trend towards larger constraint length convolutional codes. Up to now with Viterbi decoding constraint length 7 has been almost a standard, but with better technology, it's quite feasible to go up to larger constraint lengths. We are now building a constraint length 9, Viterbi decoder for example, and even up to constraint length 13 would be reasonable at lower data rates. In some of these AJ applications where you're really talking about fairly low data rates anyway, you could do decoding with higher constraint lengths

Another trend is a wider use of sequential decoding systems. In the future I expect more and more decoders to be built on chips or just a few chips With this different implementation approach sequential decoding should also be considered. It's conceivable that when the decoding is to be implemented on a chip, sequential decoding may be better than Viterbi decoding. We have built a sequential decoder on a chip and other people are building Viterbi decoders on a chip, so that is coming into being already. You have to keep an open mind when you're selecting decoding algorithms. I have a sequential decoder chip that I'll show you later and we can talk about that in more detail.

Another trend is wider use of binary decoders on non-binary channels. In particular, I'm thinking of MFSK type channels, with soft decision convolutional decoding as opposed to (say) non-binary dual K or a triple K convolutional coding, or even a hard decision type decoding algorithm. Again, these binary codes provide a greater flexibility.

What I'd like to do now is concentrate on a few areas and look at them in a little more depth. The first one I'd like to look at is an example of a decoder on a chip see Figure 5. First of all, I don't want to mislead you when I say a decoder on a chip. The guts of the decoder are on the chip but if you really want a stand alone decoder quite often it takes more than just that chip. The one I am most familiar with is the one Linkabit has implemented. It's a sequential decoder chip using a modified Fano algorithm, and I just happen to have one here. This particular chip is one that didn't work, and we've taken off the top of it. You can see the little chip in the middle; most of what I am holding here is the carrier and the pins. The chip itself is just that little dark thing in the center So this is the whole encoder-decoder chip which does not include the memory that you would have external to this. It provides for the synchronization, but it does not provide some of the clocks and things of that nature. That's why I say, if you really want a complete stand-alone decoder you really need some extra chips and that's why I say the complete decoder up here would take approximately 10 ICs. Although again, if you're including this as part of some other system, some of these things might be available already. So it's not necessarily 10 additional chips.

This particular chip handles code rates of 1/2, 3/4 or 7/8 (you just specify on the pins which decoder you want), and the various constraint lengths here go from 36 all the way up to 91. You'll recall that with sequential decoding, going to a larger constraint length doesn't really increase the implementation complexity that much. It's not like Viterbi decoding where decoding complexity approximately doubles as you increase the constraint length by 1 and of course 36 would be completely unreasonable. This particular chip uses 2 bits of quantization. The reason for using that instead of 3 bits is that it's easier to implement on a chip and you get most of a soft decision gain by

going to 2. Three buys you some additional gain, but again it depends on what modulation environment you have. You get most of the soft decision gain with 2 bit quantization.

Another characteristic of sequential decoding is that performance depends on data rate, unlike Viterbi decoding. This particular chip that we have now that we've actually tested (and we have a lot of them that we're delivering), goes up to 600 kilobits per second. We are also working on a faster version which will go up to 1.5 megabits per second. Performance improves, of course, as you go to lower data rates.

Figure 6 is an example of the performance of this chip and two other Viterbi convolutional decoding systems. I've picked the highest data rate at which this chip can operate, so it's sort of a worst case for this chip. The interference is additive white Gaussian noise and the modulation is BPSK. There is the constraint length 7 rate 1/2 Viterbi decoder, which is the standard algorithm, and we can achieve that. The other one, the constraint length 5 rate 1/2 Viterbi with 2 bit soft quantization is also being implemented on a chip (not by us) but I've put in what I think the performance is there based on theoretical considerations. All the other ones, by the way are measured performance numbers.

* QUESTION: What are those two upper numbers, why do you get better gain with the second decoder?

* ANSWER: These aren't coding gains. This is the E_b/N_0 you need. The lower numbers indicate the better performance here. The constraint lengths very from 36 to 91. You can select the code rate you want, but the constraint length is fixed in the chip for each rate, and you can't do anything about that. When I say P_b, that's error probability per information bit. Of course the error characteristics are different: sequential decoding would typically have longer bursts but P_b is the average output bit-error rate. The other thing I should point out is that even with the rate 7/8, we are doing roughly the same as with the constraint length 5 Viterbi decoding.

Switching gears again, going to a slightly different topic see Figure 7, in the future I think there's going to be more and more MFSK type systems, and we should try and concentrate on what is the best strategy for that type of a system. It looks like, and I'll try and give you some justification for this, a good approach is to use soft-decision binary convolutional codes, although this hasn't necessarily been what's been done in the past Usually when you have a non-binary channel, the first thing you think of doing is to match a code to that alphabet size, or you go to a hard decision scheme and use a binary code. By the way, when I say, "using a binary code on a non-binary channel", I'm assuming interleaving. Another way of looking at it, at least conceptually, for an octal MFSK is to have 3 separate encoder-decoders, one for each bit of the symbol. Of course in practice you wouldn't have that, you'd have interleaving. But for analysis purposes and conceptually you can look at it that way.

There are a couple of problems with binary codes on non-binary channels (in fact Jim Omura will be expanding on some of this). One is, what sort of quality measure do you use? There are many that could be used. The approach should

provide good performance for all types of environments, not just additive white Gaussian noise. The other thing we really have to look at closer is the comparison with the M-ary systems a far as best possible expected performance and complexity of decoding.

I've listed two metrics that could be used in Figure 8. Of course for the sign bit you just look at the output, pick the largest matched filter output, and that gives you the sign bit. But there are various approaches to come up with a quality measure. One approach that Andy Viterbi gave in a paper at MILCOM '82, was to take the ratio of the largest filter output to the next largest and quantize that. That gives you some indication of quality. In this approach of course you have the same quality for all of the bits of the M-ary modulation symbol. Again the thing you have to check is to make sure this is good in all kinds of environments. That's where we need a closer look at some of these quality measures

Another approach is to assign a different quality to each of the M bits of a symbol. One method is to divide the set of symbols into two groups, depending on whether they have the same bit in a specified position or not. Take the maximum filter output of each group and compare those two, the difference being a quality indicator.

* QUESTION: Are all errors equally-likely?

* ANSWER: There are various jamming strategies that you have to consider, not just the Gaussian case. You really have to look at the jamming strategies to verify this. In general, however, the incorrect symbol errors are equally likely.

* QUESTION: Why use quality measures?

* ANSWER: They improve performance. It's just like normal additive white Gaussian noise with BPSK. Going to a soft decision approach buys you something. It's not necessarily 2 dB anymore, but it does help, in fact that's what Figure 9 shows.

In Figure 9, compare the performance of several systems based on the minimum E_b/N_0 you need to operate at the computational cutoff rate. Results are given for octal and binary coding approaches with hard and soft quantization. The first thing you can see here is that there is a big improvement, or sizeable anyway, in going to soft quantization. No quantization represents an infinitely fine quantization. The binary soft decision scheme, is Andy Viterbi's ratio threshold approach and by 2 bits soft quantization I mean there's one bit of quality and one bit of sign. So that's a coarse type of quantization. And yet you can see, that you gain quite a bit just from that I quality bit. And you might guess, that if you go to another bit of quality, you would improve performance a little more. Again there are different types of environments here. When I say worst-case, partial band Gaussian jamming, I am letting the jammer concentrate his power in the fraction of the band that is best for him.

* QUESTION: What was the spectrum of the Gaussian Jammer in Figure 9?

* ANSWER: Do you mean the fraction of the band jammed? I don't remember. I think I have it in my notes, and you can

see me later. It probably is fairly narrow because there's quite a loss here relative to AWGN. If it was anywhere close to 1 I wouldn't expect this much of a difference.

I just want to point out that and there is a big difference between soft- and hard-decision decoding so you would definitely want to go to the soft-decision decoding. In fact, as I noted, this is one area where we really need more work. Especially since I think it will be one of the more common channels in the future.

Figure 10 summarizes this binary vs. non-binary code comparison. That is, binary codes are capable of, at least in some cases, performance approximately the same as non-binary codes, and have implementation advantages.

I'd like to switch again to another topic see Figure 11, and this has to do with decoding data from several sources in a TDMA application. For example, if you have a downlink decoder that has to handle data from many different separately encoded sources, there are various ways of doing this. You can have a separate decoder for each source bit that's a little wasteful. You could have many users, so you certainly wouldn't want to do that Another approach would be to share a basic Viterbi decoder with the metric and path histories switched in and out. Whenever you change channels and decode from a different source, you can take these metrics and path histories, put them off in some memory, grab the new ones and decode for a little bit, and then throw them back and bring in the next. We are building a decoder that does that for up to 16 separately encoded sources.

Another approach is to actually form a block code somehow. Then you could decode one block of data before you go on to a block from another source Then the question is, "How do you form those blocks in a convolutional coding system? One possibility is to put in so-called tail or flush bits, a sequence of all zeros, and that would terminate the code and make it into a block code. The problem here is that there will be some overhead. That tail will cost you a little bit, and you will incur, some overhead loss. I've noted that this is usually a very good approach, especially if you are designing the system from scratch; just make your block length long enough so that the overhead loss is not a problem. Unfortunately, if the system is already there and you are trying to retrofit it, you sometimes don't have that flexibility. Another scheme that was proposed a few years ago, is so-called tail-biting. The basic idea here is that there is a wrap-around effect where you don't have to send this added tail. Unfortunately you don't get that for free. You don't have the tail loss to contend with but there is a penalty, and that penalty is typically in speed. At the decoder you have to decode several times faster typically. So again, forming blocks doesn't come for free, and also there could be a performance loss if you are not careful. So you really have to look at performance close.

Just to give you a little more on what these tail-biting codes are, they are a type of code where you don't have to send a tail to terminate the code to turn it into a block code If you have a sequence of N input bits, normally you just start shifting these into some encoder register that's filled with zeros initially, shift one bit in and get V out and keep doing that. The way to get around sending the tail is to initialize the encoder using the last K-1 bits to be sent, and when you shift them in, you don't start sending output bits until after these K-1 bits are fully loaded A feature of these codes is that they start out in some state, and end up in the same state. At the receiving site, you don't know

what that initial state is, but it's the same as the final state. In fact, that's the property that you use to do the decoding.

Figure 13 is a simple example of the basic idea behind the decoding of the tail-biting code. I'm thinking mainly of a Viterbi decoder here, but you could extend this to sequential or feedback or other types of decoding as well. This example is a constraint length 3 code with 4 encoder states, and this is my version of a trellis. There are 4 initial states on the left and 4 final states on the right. When you start out, you don't know what the initial state is, so you have paths coming from each possible state. After you've decoded a while, hopefully, these paths merge. Then at the end, of course, they diverge again to each of the final states. If you tried to estimate that sequence just based on that single path, you would have unreliable data at the ends, but in the middle, since everything is merged, it would be reliable. This makes you think if you could just extend this somehow and use only the middle part, that might be a good approach. Fortunately since the first and last states are the same, you can in effect, make this look like a longer received sequence by just repeating the received sequence several times. I've shown two passes in Figure 13 but you might want to do it even more than two. If you decoded the same sequence twice, the paths would diverge at both ends but hopefully in the middle they would be merged for a block length and this merged portion of the trellis could be used to estimate the data sequence.

Figure 14 is an example of this. Tail-biting codes are really block codes, and you can describe one with a generator matrix. All the words would be linear combinations of the rows of the generator matrix. This example is for a constraint length 9, rate 1/2 code. This is the best convolutional code with these parameters. This tail-biting code is a (24,12) code, and I just want to point out that we know that the extended Golay code has a distance 8. This, as a convolutional code, has a free distance of 12, so you must expect some loss. The interesting thing is that if you look at the properties of this particular code, it is exactly equivalent to the extended Golay code. All the distance properties for all code words match up. As a class of codes, these tail-biting codes do include some good codes. This one, if nothing else, also this code equivalence makes it possible to do soft-decision decoding for the extended Golay code using a convolutional decoder designed for the optimum K=9, R=1/2 code. These results are summarized in Figure 15.

There are other interesting properties of tail-biting codes that we don't have time to go into, e.g., there seems to be a mapping of bad convolutional codes into tail-biting bad codes. Catastrophic convolutional codes seem to map into catastrophic or worthless block codes where two inputs lead to the same output; a bad situation for sure. It looks like this may be true in general but I haven't looked at that. The last thing I'd like to point out is that even though I initially brought these codes up as a way of handling short bursts, the best payoff in developing a better understanding of this very structured class of codes is that it could lead to a better understanding of convolutional codes in general. I think this will be the better payoff in the long run.

SYSTEM NEEDS

- IMPROVED PERFORMANCE IN
  - AWGN
  - JAMMING ENVIRONMENTS
  - SCINTILLATION (NUCLEAR AND ATMOSPHERIC)
  - MULTIPATH
  - END-TO-END SATELLITE CODING SYSTEMS WHERE THE UPLINK AND THE DOWNLINK INTERFERENCE CONTRIBUTE SIGNIFICANTLY TO THE ERROR RATE.
- SMALLER SIZE, WEIGHT AND POWER DECODERS.
- LOWER COST.

Figure 1

SYSTEM NEEDS (CONT.)

- IMPROVED FLEXIBILITY
  - GOOD PERFORMANCE WITH VARIOUS MODULATIONS (DPSK AND MFSK WITH VARIOUS NUMBERS OF CHIP REPETITIONS WILL BE MOST COMMON IN MILITARY SYSTEMS).
  - GOOD PERFORMANCE IN THE PRESENCE OF INTELLIGENT JAMMERS: SYSTEM SHOULD NOT BE VULNERABLE TO SPECIAL JAMMING STRATEGIES THAT USE MEASURED CHANNEL INFORMATION (AGC, PRESENCE OF A JAMMER, ETC.).
- BETTER APPROACH TO ACCOMMODATE THE DECODING OF DATA FROM SEPARATELY ENCODED SOURCES.
- CAPABILITY OF ACCOMMODATING HIGHER DATA RATES ESPECIALLY IN COMMERCIAL SYSTEMS.
- FASTER DECODER NODE SYNC AND, IF NECESSARY, PHASE-AMBIGUITY RESOLUTION.
- FASTER SYNCHRONIZATION AND MINIMUM THROUGHPUT DELAY FOR SYSTEMS WITH INTERLEAVER/DEINTERLEAVERS.

Figure 2

LONG-TERM TRENDS

- LARGER CONSTRAINT LEGNTH (K) VITERBI DECODED CONVOLUTIONAL CODES WILL BECOME COMMON.
    - MOST OF THE RATE R=1/2 SYSTEMS TODAY USE K=7.
    - K=9, R=1/2 DECODERS ARE NOW BEING BUILT AND K=13 IS REASONABLE FOR SOME LOW DATA RATE APPLICATIONS.
    - A K=9, R=1/2 SYSTEM IS .6 dB SUPERIOR TO A K=7 SYSTEM WITH BPSK/QPSK MODULATION AND AWGN AT $P_b=10^{-5}$.
- MUCH WIDER USE OF SEQUENTIAL DECODING SYSTEMS.
    - SEQUENTIAL DECODING SYSTEMS ARE IN MANY CASES CAPABLE OF ACHIEVING LARGER CODING GAINS THAN VITERBI DECODERS OF A SIMILAR IMPLEMENTATION COMPLEXITY.
    - SEQUENTIAL DECODING IS EASILY AMENABLE TO IMPLEMENTATION ON A LSI CHIP.

Figure 3

LONG-TERM TRENDS (CONT.)

- WIDER USE OF DECODERS IMPLEMENTED ON A FEW, OR SINGLE, CHIPS.
- WIDER USE OF BINARY CODES WITH INTERLEAVING ON NONBINARY CHANNELS.
    - PROVIDES A GREATER GAIN THAN NONBINARY (E.G., DUAL-K, TRIPLE-K) CODES FOR A GIVEN IMPLEMENTATION COMPLEXITY.
    - PROVIDES GREATER FLEXIBILITY THAN NONBINARY CODES WITH DIFFERENT MODULATIONS.

Figure 4

EXAMPLE OF NEXT GENERATION DECODERS: SEQUENTIAL DECODER CHIP

- LINKABIT HAS DEVELOPED A CONVOLUTIONAL ENCODER/FANO ALGORITHM SEQUENTIAL DECODER ON A nMOS CHIP.
- TOTAL DECODER TAKES 10 IC's.
- MODES ARE: R=1/2, K=36; R=3/4, K=63; R=7/8, K=91.
- 2-BIT SOFT DECISIONS.
- MAXIMUM DATA RATE - 600 K BPS WITH A 1.544 M BPS VERSION UNDER DEVELOPMENT.
- PERFORMANCE DEGRADES WITH INCREASING DATA RATES.

Figure 5

PERFORMANCE SUMMARY

TYPE OF DECODER	E_b/N_o IN dB FOR $P_b = 10^{-5}$ *
R = 1/2 LSI SEQUENTIAL	4.5
R = 3/4 LSI SEQUENTIAL	5.2
R = 7/8 LSI SEQUENTIAL	5.8
K = 5, R = 1/2 VITERBI (2-BIT SOFT)	5.7
K = 7, R = 1/2 VITERBI (3-BIT SOFT)	4.5

*BPSK MODULATION, AWGN AND 600 K BPS DATA RATE.

Figure 6

EXAMPLE OF NEXT GENERATION DECODERS: SEQUENTIAL DECODER CHIP

- LINKABIT HAS DEVELOPED A CONVOLUTIONAL ENCODER/FANO ALGORITHM SEQUENTIAL DECODER ON A nMOS CHIP.
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R = 7/8 LSI SEQUENTIAL	5.8
K = 5, R = 1/2 VITERBI (2-BIT SOFT)	5.7
K = 7, R = 1/2 VITERBI (3-BIT SOFT)	4.5

*BPSK MODULATION AWGN AND 600 K BPS DATA RATE.

Figure 6

OPTIMIZING THE PERFORMANCE OF SOFT-DECISION BINARY CODING SYSTEMS ON NONBINARY CHANNELS

- THIS IS THE MOST NOTABLE AREA WHERE MORE WORK IS NEEDED.
- MOST ANALYSIS WORK TO DATE HAS BEEN FOR BINARY CODES ON BINARY CHANNELS OR FOR NONBINARY CODES ON NONBINARY CHANNELS.
- IN THE FUTURE THE USE OF BINARY SOFT-DECISION DECODING ON CHANNELS WITH MFSK MODULATION WILL BE INCREASINGLY COMMON.
- WHAT IS A GOOD QUALITY MEASURE FOR SUCH DECODING IN VARIOUS ENVIRONMENTS?
- HOW DOES THE PERFORMANCE OF BINARY CODES COMPARE WITH THAT OF M-ARY CODES ON M-ARY CHANNELS?

Figure 7

QUALITY MEASURES FOR BINARY CODES ON 2^M-ARY MFSK CHANNELS

- THE M SIGN BITS ARE DETERMINED FROM THE I.D. OF THE LARGEST MATCHED FILTER OUTPUT.
- SEVERAL APPROACHES FOR DETERMINING QUALITY BITS ARE POSSIBLE.
  - APPROACH 1 - ASSIGN THE SAME QUALITY TO ALL M BITS OF A SYMBOL AND BASE IT ON A QUANTIZATION OF THE RATIO OF THE LARGEST TO THE NEXT LARGEST FILTER OUTPUTS.
  - APPROACH 2 - DETERMINE THE QUALITY BITS ASSOCIATED WITH EACH BIT OF A SYMBOL SEPARATELY. DIVIDE THE FILTER OUTPUTS INTO 2 GROUPS BASED ON THE SIGN OF THE BIT IN QUESTION. THEN ASSIGN QUALITY BITS BASED ON A QUANTIZATION OF THE DIFFERENCE OF THE MAXIMUM FILTER OUTPUTS IN EACH GROUP.

Figure 8

PERFORMANCE OF BINARY AND OCTAL CODES WITH OCTAL-MFSK MODULATION AND R = 1/2 CODING

CHANNEL	MINIMUM E_b/N_o IN dB FOR $R < R_o$			
	OCTAL CODE		BINARY CODE	
	Hard Quantization	No Quantization	Hard Quantization	Ratio-Threshold 2-bit Soft Quantization
AWGN	6.56	4.87	6.47	5.7
WORST-CASE PARTIAL BAND GAUSSIAN JAMMING	9.66		9.40	6.4
RAYLEIGH FADING	13.35	9.12	13.07	9.8

Figure 9

PRELIMINARY BINARY VS. NONBINARY CODING CONCLUSIONS FOR MFSK SYSTEM

- BINARY CODES ARE CAPABLE OF APPROXIMATELY THE SAME, AND IN SOME CASES BETTER, PERFORMANCE THAN NONBINARY CODES.
- SINCE BINARY ENCODER/DECODERS ARE EASIER TO IMPLEMENT THAN NONBINARY ENCODER/DECODERS, BINARY CODES ARE A GOOD CHOICE.

Figure 10

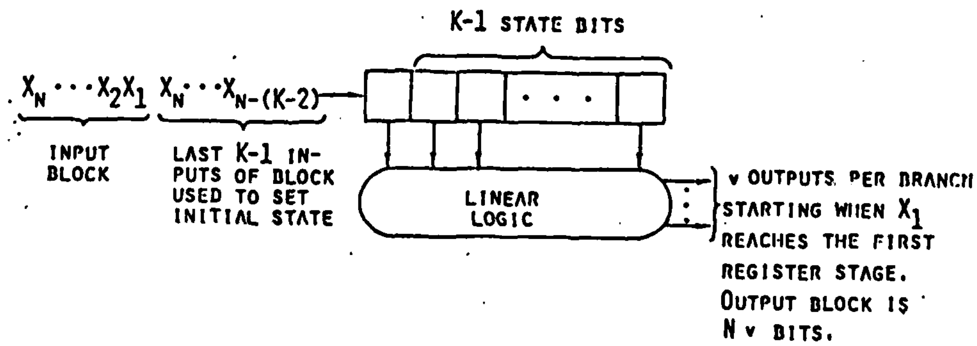
DECODING OF DATA FROM MULTIPLE SOURCES (E.G., TDMA)

- SEPARATE DECODERS CAN BE USED FOR THE DATA FROM DIFFERENT ENCODING SOURCES.
- A SINGLE SHARED DECODER WITH SWITCHING OF METRICS AND PATH MEMORIES CAN BE USED.
- DATA CAN BE GROUPED INTO BLOCKS AND A SINGLE SHARED DECODER USED.
  - TAIL (OR FLUSH) BITS CAN BE ADDED TO TERMINATE THE CONVOLUTIONAL ENCODED SEQUENCES.
    - LOSS DUE TO THE NEED FOR OVERHEAD BITS TRANSMITTED.
    - A GOOD APPROACH IF BLOCKS ARE LONG ENOUGH.
  - TAIL BITING ENCODING/DECODING CAN BE USED.
    - NO OVERHEAD BITS ARE REQUIRED.
    - DECODER MUST OPERATE AT A HIGHER RATE THAN IF TAIL BITING IS NOT USED.
    - PERFORMANCE OF PARTICULAR SYSTEM MUST BE DETERMINED.

Figure 11

TAIL BITING FEATURES

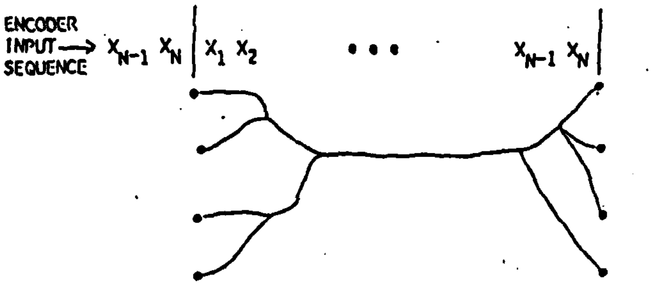
- TAIL BITING IS A PROCEDURE WHEREBY SHORT BLOCKS OF DATA CAN BE ENCODED WITHOUT THE NEED TO TAIL OFF EACH BLOCK.
- TAIL BITING EXAMPLE FOR AN $R = 1/v$ CODE



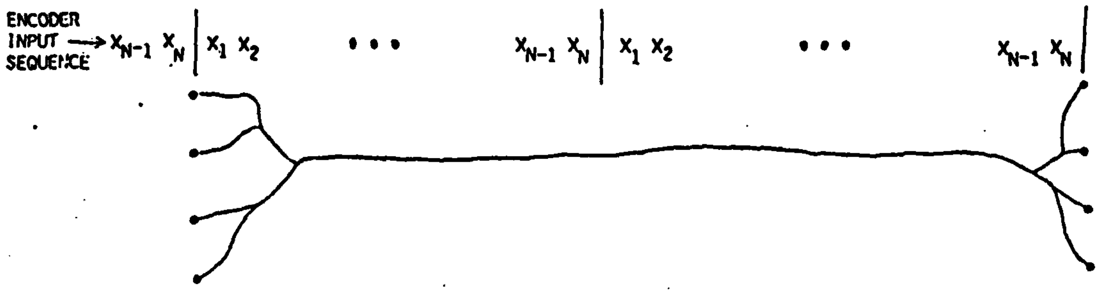
- NOTE THAT THE INITIAL AND THE FINAL STATES IN THE TRELLIS PATH TRACED BY A TAIL BITING CODEWORD ARE THE SAME. THIS MAKES IT POSSIBLE TO TREAT THE RECEIVED SEQUENCE AS A LONG SEQUENCE COMPOSED OF SEVERAL REPETITIONS OF THE RECEIVED BLOCK AND TO SELECT A DECODER OUTPUT SEQUENCE FROM THE MORE RELIABLE CENTER PORTION OF THE TRELLIS.

Figure 12

EXAMPLE OF TAIL BITING DECODING FOR A K = 3 CODE



A) TRELLIS WITH ONE DECODING PASS.



B) TRELLIS WITH TWO DECODING PASSES. MERGED PATHS IN THE CENTER PORTION OF THE TRELLIS PROVIDE A RELIABLE ESTIMATE OF THE DATA BLOCK.

Figure 13

EXAMPLE OF A K = 9, R = 1/2, BLOCK LENGTH 24 TAIL BITING CONVOLUTIONAL CODE

$$G = \begin{bmatrix}
1&1&1&0&1&1&1&1&0&1&1&0&0&0&1&0&1&1&&&&&& \\
&&1&1&1&0&1&1&1&1&0&1&1&0&0&0&1&0&1&1&&&& \\
&&&&1&1&1&0&1&1&1&1&0&1&1&0&0&0&1&0&1&1&& \\
&&&&&&1&1&1&0&1&1&1&1&0&1&1&0&0&0&1&0&1&1 \\
1&1&&&&&&&1&1&1&0&1&1&1&1&0&1&1&0&0&0&1&0 \\
1&0&1&1&&&&&&&1&1&1&0&1&1&1&1&0&1&1&0&0&0 \\
0&0&1&0&1&1&&&&&&&1&1&1&0&1&1&1&1&0&1&1&0 \\
1&0&0&0&1&0&1&1&&&&&&&1&1&1&0&1&1&1&1&0&1 \\
0&1&1&0&0&0&1&0&1&1&&&&&&&1&1&1&0&1&1&1&1 \\
1&1&0&1&1&0&0&0&1&0&1&1&&&&&&&1&1&1&0&1&1 \\
1&1&1&1&0&1&1&0&0&0&1&0&1&1&&&&&&&1&1&1&0 \\
1&0&1&1&1&1&0&1&1&0&0&0&1&0&1&1&&&&&&&1&1
\end{bmatrix}$$

- CODE GENERATORS ARE FOR THE BEST K=9, R=1/2 CODE GIVEN BY 1 1 1 1 0 1 0 1 1
  1 0 1 1 1 0 0 0 1.

- IN SOME CASES, CAREFULLY SELECTED TAIL BITING CONVOLUTIONAL CODING SYSTEMS HAVE VIRTUALLY THE SAME BIT ERROR RATE PERFORMANCE AS THAT OF THE BASIC CONVOLUTIONAL CODING SYSTEM.

Figure 14

EXAMPLE OF A K = 9, R = 1/2, BLOCK LENGTH 24 TAIL BITING CONVOLUTIONAL CODE (CONT.)

- FOR THE EXAMPLE GIVEN HERE, THE BASIC CONVOLUTIONAL CODE HAS d_f = 12 BUT THE (24, 12) TAIL BITING CODE HAS d = 8. SO THERE WILL BE A PERFORMANCE LOSS IN THIS CASE.
  - FOR THIS EXAMPLE THE PERFORMANCE DEGRADATION ON A BPSK/QPSK AWGN CHANNEL AT SMALL ERROR RATES IS APPROXIMATELY 10 LOG (12/8) = 1.7 DB.
  - PRACTICALLY THE BLOCK LENGTH SHOULD BE CONSIDERABLY LONGER THAN 3 CONSTRAINT LENGTHS TO AVOID DEGRADATIONS DUE TO TAIL BITING.
- AN INTERESTING THEORETICAL PROPERTY OF THIS TAIL BITING CODE IS THAT:

> THE (24, 12) TAIL BITING CONVOLUTIONAL CODE FORMED FROM THE OPTIMUM K = 9, R = 1/2 CONVOLUTIONAL CODE IS EQUIVALENT (SAME DISTANCE PROPERTIES) TO THE EXTENDED GOLAY CODE.

Figure 15

SIDE INFORMATION AND CODING METRICS IN JAMMING CHANNELS[1]

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In a coded communication system the optimum metric used for decoder decisions is the maximum likelihood metric which generally assumes knowledge of the channel statistics. For many real channels, particularly a jamming channel, there is incomplete knowledge of the channel statistics and non-optimum metrics must be used. Generally in these cases "robust metrics" are desirable. By robust metrics we mean decoder decision metrics that yield coded bit error rates "close" to a baseline broadband noise jammer case regardless of the true waveform used by the jammer. In addition side information about the jamming channel may be incorporated in a metric.

Using a generalized cutoff rate parameter as well as specific codes, we compare various metrics and the impact of side information on coded antijam communication systems. In particular we examine the sensitivity of the coded bit error probability to the side information and metric used as a function of the coded symbol energy-to-equivalent jammer noise ratio. The spread spectrum signals we use in our examples are the coherent direct sequence spread binary phase shift keying (DS/BPSK) and the noncoherent frequency hopped M-ary frequency shift keying (FH/MFSK) modulations. Side information include constant channel parameters and instantaneous jammer states.

Details of this presentation in order of appearance are

* General analysis of coded communication systems with general metrics

* Generalized cutoff rate parameter.

  - Usual hard and soft decision metrics for the AWGN channel

  - Coding bounds and cutoff rates

DS/BPSK with pulse jamming.

  - Jammer state information

  - Hard and soft decision metrics

  - Repeat code

  - Jammer state estimates (El-Wailly and Costello)[8]

FH/MFSK with partial band jamming.

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[1] This work was supported by NAVAIR through NRL Contract Award No. NOOO14-80-K-0935

- Jammer state information

- Hard and soft decision metrics

- Repeat code/diversity

- Two bit metric (Viterbi)[9]

- List metric (Crepeau, Creighton and Omura)[9]

Key References

1. Viterbi and Jacobs, "Advances in Coding and Modulation for Noncoherent Channels Affected by Fading, Partial Band, and Multiple-Access Interference," chapter in Advances in Communication Systems, edited by A.V. Balakrishnan, Academic Press, 1975.

2. Viterbi, "Spread Spectrum Communications - Myths and Realities," IEEE Comm. Soc. Mag., Vol. 17, May 1979, pp. 11-18.

3. Omura and Levitt, "A General Analysis of Anti-Jam Communication Systems," NTC '81, New Orleans, November 29-December 3, 1981.

4. Omura and Levitt, "Coding Tradeoff for Improved Performance of FH/MFSK Systems in Partial Band Noise," NTC '81, New Orleans, November 29-December 3, 1981.

5. Omura and Levitt, "Coded Error Probability Evaluation for Antijam Communication," IEEE Trans. on Comm., Vol. COM-30, No. 5, May 1982, pp. 896-904.

6. Viterbi, "A Nonparametric Mitigation Technique for Partial Band Jamming of MFSK Frequency Hopped Communication Links," MILCOM '82.

7. Omura and Avidor, "Analysis of FH/MFSK Systems in Non-Uniform Rayleigh Fading Channels," MILCOM '82, Boston, MA, October 17-20, 1982

8. El-Wailly and Costello, "Performance Analysis of Coded D/S Spread Spectrum A/J Receivers with Jammer State Estimates," GLOBECOM '82, Houston, TX.

9. Crepeau, Creighton and Omura, "Performance of FH/MFSK with List Metric Against Partial Band Noise Jamming," submitted to MILCOM '83.

10. Simon, Omura, Scholtz, and Levitt, A UNIFIED APPROACH TO SPREAD SPECTRUM COMMUNICATION, Book to appear in 1983.

WELCH: Our last speaker, Jim Omura, who received his Ph.D. at Stanford, is now a professor at UCLA. He is co-author of the forthcoming book A Unified Approach to Spread Spectrum Communications along with Levitt, Scholtz, and Simon, all of whom are here. Jim does not belong to a thriving company. (Laughter)

OMURA: (With great regret!!) Actually, I don't have anything new to present. In the process of writing a book on spread spectrum, what has happened is that I've been forced to put together, in some coherent manner, a lot of the known results in the spread-spectrum anti-jamming area And so what I'm going to talk about with you today, is just more or less my organization of the subject, especially how coding and modulation interplay. The results are not necessarily new but I'm just trying to present it in a coherent manner, motivated by writing this text book. From this, of course, some new questions may come up and I hope that this talk will be a little more open-ended in pointing out some questions or giving a different way of looking at things.

My particular section of the book has to do with side information and how does one use side information in anti-jam communications. This also involves multiple-access situations. Here we move away from the classical additive white Gaussian noise channel and raise other issues in terms of a good receiver design. You may or may not know what is a maximum-likelihood receiver, and so forth. I'll illustrate that with some simple examples, again well-known, but they lead into some new questions.

For instance (see Figure 1), classical modulation/demodulation theory generally assumes a white Gaussian noise channel Here we have modulation and the optimum receiver. If it is coherent, this is, of course the usual bank of correlators You might have a bank of matched filters if it is a multiple-access scheme. If this happens to be MFSK then your optimum receiver in the white Gaussian noise case is going to be a bunch of energy detectors and you are going to look at the energies associated with each of the frequencies. That's the modulation end of things and spread-spectrum can be included in there.

Next we have the coding end in Figure 1, where from a coding theorist's point of view, generally you worry about sending sequences of symbols into a channel (and usually you assume the channel is some kind of discrete memoryless channel), you get an output sequence and again you talk about optimum decoding, typically some sort of maximum likelihood decoding algorithm. I'm interested in discussing the interface between these two. Namely, how do you take bits that are going into this coded system (the real channel is actually embedded in the DMC somewhere and there's an interface between the coded symbols and the actual bits going into the modulator and out of the demodulator and back into symbols that then go on) and go through some sort of decoder at the receiver end. So the subject really is that interface between modulation and coding that I want to spend a little time discussing today.

Typically most of the coding schemes are designed to work well in the memoryless channel. And so if the channel has memory, you're going to have, as Elwyn and Joe discussed, interleavers/deinterleavers This interleaver/deinterleaver might be part of this interface scheme So the question is, how does one take the output of the demodulator (energy detector in the MFSK case) and somehow quantize that appropriately so that then you can go through a decoding process, and hopefully simplify the decoding process. In addition you have the problem that you don't really

have an additive white Gaussian noise channel; you may have a much more complicated channel where you have unknown channel parameters, fading, jamming, other signals in a multiple-access environment, etc., so what was maximum likelihood for white Gaussian noise channel no longer holds true in this more general setting. In fact, most of the time you don't even know what the channel statistics are so another question concerns what metrics should you use. Generally, those in the spread spectrum world know that you want a robust metric, something that is effective for all kinds of things that can go in the channel. The question is, "How do you start to evaluate various robust metrics and compare them?"

Figure 2 is the classical example of a receiver interface, perhaps the most trivial of all. You take binary phase-shift keying and one of the first things you do, if you have coding, is to quantize it. There is a difference having to do with how you quantize the correlator outputs. Typically if you do a hard decision, then you wind up with the classical binary symmetric channel. On the other hand, if you were just to take the real output (soft decision) rather than quantize it, there's a difference in performance with coding. Most of you are of course familiar with the well-known constraint length 7, rate 1/2 convolutional code as it functions in a white Gaussian noise environment. You see in Figure 3 that there is a roughly a 2 dB difference between hard and soft decision (that's well known). The approach that I've been taking in this book is to look at the cutoff rates for these channels, because if you look at the cutoff rate, for instance in Figure 4 for this example, you notice that the same 2 dB difference shows up in the cutoff rate. That same 2 dB difference also appears for all binary convolutional codes, so that looking at the cutoff rate difference is like looking at a performance difference that seems to hold true regardless of what code you use. Therefore the cutoff rate seems like a natural way to evaluate what kind of quantization, or what kind of metric, one might use in the channel.

In addition to the kind of quantization that one uses in the channel (hard decision/soft decision), or various levels of quantization, there's another issue that comes up, and that has to do with what kind of decoding metric one should use. Once you get an equivalent coding channel, the maximum likelihood metric assumes you know the conditional probabilities of that coding channel, but in fact oftentimes you don't. You'd like to use maximum likelihood but you can't since you may not know those channel probabilities.

Just to illustrate where you have some problems here, let's again consider coded binary phase-shift keying (Figure 5). We are going to take a trivial code like the repeat code where we just repeat the symbol "zero" N times, or repeat the symbol "one" N times (that's diversity, or the trivial code). Now if you had a maximum likelihood receiver, and the noise was just white Gaussian and the noise variance doesn't change, you'd have the usual correlator receiver. That's the best you can do. The metric on a per bit basis or per symbol basis would just be straight correlation. Now suppose the noise variance in the channel (this is just a made-up example, still using Gaussian noise), changes as a function of time Here the maximum likelihood thing to do, is correlation, but weighted appropriately by the noise variance. Now if, channel conditions are changing, the noise variance would be side information. You'd have to get that information in addition to running your system. The question is, is it worthwhile doing that? If you don't do that, you would be doing something sub-optimally. How well does that perform,

when in fact the channel noise variance is changing in time?

These are of course well-known issues that people have looked at in the past. I want to illustrate, in a jamming context, where this occurs (see Figure 6). In a direct sequence spread spectrum system, if you have a pulse jammer which is on a fraction of the time ρ, and off a fraction of a time 1ρ, what happens is you have virtually very little noise, or no noise for some of the time and lots of noise at other times. Now if you knew the noise variance that each symbol encounters when going across the channel, you would of course use the maximum likelihood rule. If you don't know it, you might be tempted to just use the general rule you use in the white Gaussian noise channel. Now for the uncoded case, if you used the AWGN rule in the fraction ρ of time that the jammer on, and now the jammer happens to know your system and choose the worst choice ρ^* of ρ, then the well-known result is that the jammer essentially hurts you by the amount shown in Figure 7. The performance of uncorrelated binary phase-shift keying with a continuous jammer ($\rho=1$) as opposed to a jammer picking up the worst worst pulse duration is also given in Figure 7. There is about a 30 dB difference or a 34 dB difference at the lower end of these curves. What's happening is not surprising since the $\rho=1$ curve for error probability is a very sensitive function of the SNR. Obviously if the jammer has the ability to move the SNR slightly, he is going to cause a tremendous change in the error probability so even though he's making the SNR smaller for a small fraction ρ of the time, the average is still coming way up. The effect is just like a fading channel in which the SNR changes, and obviously, just as in the fading channel, P_b varies as the reciprocal of the SNR.

In a coded system, let's look at the same problem where the jammer may be on or off (see Figure 8). When the jammer state z_k is unknown, (i.e. the variance of the noise is unknown) then I'm just going to use a conventional correlation receiver. I could use the $u_k x_k$ metric in a high precision (soft decision) processor (that's the optimum metric if the AWGN channel is not changing) or I could use the $y_k x_k$ (hard decision) metric. On the other hand, if I had side information z_k, i.e., I could measure the times when I'm being jammed (let's say in principle I could do that), I would obviously weight it by $c(z_k)$ and would actually divide by the variance of the noise and the soft-decision weighted metric would be the optimum strategy.

If we were to look at these cases for a simple code, namely m-fold repeating in time, (this is just a trivial diversity case) we can get exact expressions shown in Figure 9 for all four of these cases (soft decision/hard decision, with/without jammer state knowledge). If you were to use a soft decision case but without jammer state information (see Figure 10) and the fraction of the time the jammer jams you is ρ, you see that the error probability, even in a coded system like this diversity system, gets worse for lower values of ρ at higher values of E_b/N_0. So obviously you don't want to use the simple soft decision. In fact Figure 11 is an example where without jammer state information (JSI), but using hard decision, you do better than soft decision Clearly the soft decision metric is the wrong one to use when the jammer is pulsed and JSI is not available. This is because a pulse jammer can only do so much damage in a hard decision case. (You clip everything.)

If you were to take into account jammer state information, the whole system improves correspondingly (see Figure 12). But then this requires more side information, mainly the information that you are being jammed or you are not

being jammed on a symbol by symbol basis. That may be hard to obtain, so the tendency in this case might be to go to the hard decision case (see Figure 13) which is more robust, or to do some sort of clipping of the energy levels. These ideas have been around, but I'm just trying to formalize them here. Clearly you need some form of coding. You have tremendous coding gains against jammers, in a spread spectrum environment, and also in fading environments as well.

The general encoder and decoder and the basic structure we look at is shown in Figure 14. We assume the channel is made memoryless by interleaving/deinterleaving, and this is part of the interfacing. In the channel you might have noise, jamming, interference from other users, and so forth, a lot of which you don't know. You go through the usual despreader demodulator and here you wind up inside this (dashed) area with some sort of equivalent discrete memoryless channel and in addition you may or may not have side information available. For instance in HF frequencies you typically run sounders across the band to measure the propagation condition. That may be a kind of long term measurement that's available. On a short term basis you may determine when you're not being jammed on a symbol by symbol basis. Maybe that information is available, maybe not, and the attempt here is to structure all these possibilities in one analytical formulation. By the way, here too, the side information comes into play here at the decoder, affecting the kind of metric one might use. Even though you may have a discrete memoryless channel within the dotted line, you may not know all the conditions in the channel and so you cannot use the maximum likelihood metric. You probably are looking for a more general robust one and it may be that you have some side information that might help you improve your system

This formulation (Figure 14) is nice in the sense that, if you believe in the cutoff rate and the union-Chernoff type bounds, even without a maximum likelihood metric, you can generalize the notion of the cutoff rate using Chernoff bounds and so on. The overall bit error probability P_b for this system is in general bounded by some function of the cutoff rate (see Figure 15). And the cutoff rate R_0 is only dependent on what's inside the box and the metric. The cutoff rate R_0 generally is a function of the signal carrier or coded symbol to equivalent noise ratio, whereas the bit error probability itself also depends on the code used. So with R_0 there's a separation of what goes on in the physical channel with all this mess you might have in there, and the metric one chooses, and the specific code one is using here Hence if you want to evaluate different metrics under different channel conditions you can compare the cutoff rates (that's a code invariant comparison), and then on top of that you get your usual coding gain. The effect of coding is usually analyzed in the following way. The bit error probability is related to the symbol error probability by the code rate in bits per coded symbol

You can compute the cutoff rate (for instance) in the 4 simple examples I just cited, the binary phase-shift-keying channel with the pulse jammer in there And you can look at having jammer state information and not having jammer state information, hard decision/soft decision. Again you can see that there is a one to one correspondence of how the error probabilities differ and how the cutoff rates vary as a function of different pulse duration or pulse time (see Figures 16 and 17).

I want to briefly mention some work that is based upon this kind of formulation. One of the problems in the pulse jamming case is, "If you wanted jammer state information, how do you get it?" An

approach that was suggested by Costello and his student (see Figure 18) is to assume you have side information concerning when you are being jammed by making an estimate of it, i.e., assume you are being jammed when the signal correlator output deviates far from the signal levels that you're expecting. This puts a threshold on both sides of the expected correlator outputs, and if the next correlator measurement falls outside these thresholds, we assume that the jammer is on. And if it falls inside, assume the jammer was off, and define your metric accordingly This is just one attempt to incorporate estimates of the side information and there's been some analysis, reported in GLOBECOM 82, comparing performance for this non-ideal case with having ideal jammer state information or none at all.

For frequency hopped MFSK systems you have similar kinds of results with partial jamming and multi-tone jamming (see Figure 19). Frequency-hopped binary phase-shift-keying performance for broadband jamming is much better than for a jammer that jams a fraction of the band. You have a dual situation where, if the jammer chooses the worst fraction, poor performance results, and you get similar results from multi-tone jamming.

The MFSK problem is more complicated because now you're getting M tones out and there are actually much more options available as to what kind of metric one might choose in this environment. I'd like to just talk about some metrics that people have been looking at. One thing you must do, if you have a whole sequence of these M energy detector outputs corresponding to a coded sequence, (these are chips, you are going to combine the chips together), is to form a metric for decoding. Again the same kind of problems occur here as did in the direct sequence case with pulse jamming: Should you take weighted sums of chip energies, or should you make hard decisions, and so forth.

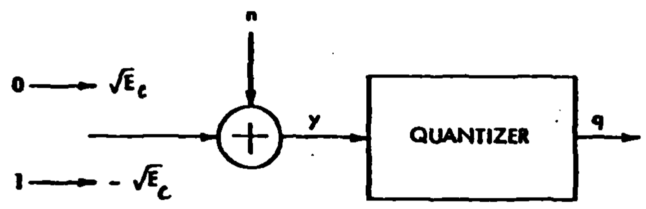
One thing you can do is make a hard decision in this FH/MFSK case. Each time M chips come out, make a hard decision as to what chip had the largest energy, assume that was the transmitted chip, and do that on a chip-by-chip basis for successive hops and put it all together and make a final decoding decision. Viterbi suggested adding one more additional bit of information to this (see Figure 20), and this is a scheme that Joe Odenwalder talked about. He says, find the largest chip energy output E^* and then for parameter A, determine if this largest one is greater than A times the next largest one, and if so, you assume that the quality is pretty good. If it's not, you assume the quality is not so good. Associated with each of the consecutive outputs is a hard decision plus a quality factor of good or bad, the additional bit. Now if you were to break this down into a binary symmetric channel, it boils down to binary symmetric channel where each bit has a hard decision bit plus a quality factor decision, and you can use Linkabit's sequential decoder on that

Here's another way of getting from an MFSK channel to bits. This one is where each 8-ary symbol has associated with it binary symbols. Suppose we are interested in the first bit. Associated with that bit are the corresponding energy terms which we use to find a metric for this bit. This metric turns out to be very close to optimum in a Rayleigh fading channel with additive white Gaussian noise. The purpose of this metric is to make a decision as to the first bit given the 8-ary channel output. You can just take the largest of the energy components here minus the energy components here and use that as though that's your correlator output which you can quantize

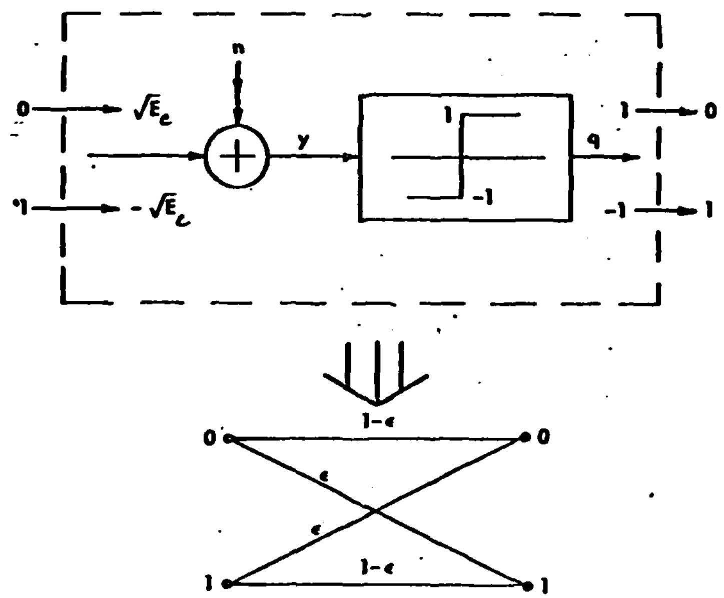
to 1 bit, 2 bit, 3 bit, (however many you want) and feed that into your usual binary convolutional decoder.

Another metric which I've been looking at is the list metric (see Figure 21 and 22). Ken Jordan used this many years ago. A list metric is not very good for the white Gaussian noise channel but it seems to have some nice properties in a more cluttered environment. The list metric looks at the energy detector outputs and rank-orders them. When you are in an environment where your intended signal may be the second or third on the list, this metric allows you to take advantage of that information and not necessarily erase the whole thing. You may have side information which allows you to take advantage of the list information.

Figure 2.



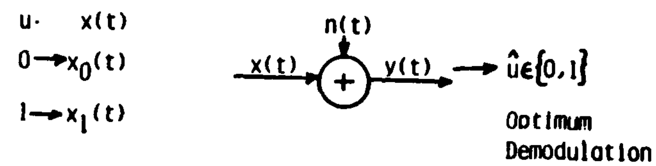
(a) GENERAL BPSK CODING CHANNEL



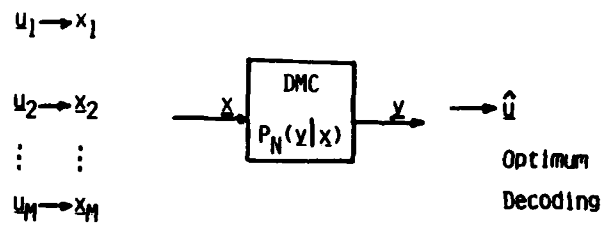
(b) HARD DECISION BPSK CODING CHANNEL

Figure 3. BPSK coding channels

Modulation/Demodulation



Encoding/Decoding



Real Coded Channels:

- Unknown Channel Parameters and/or Statistics (Fading, Jamming, CDMA)
- Interleaver Memory Limit

Figure 1

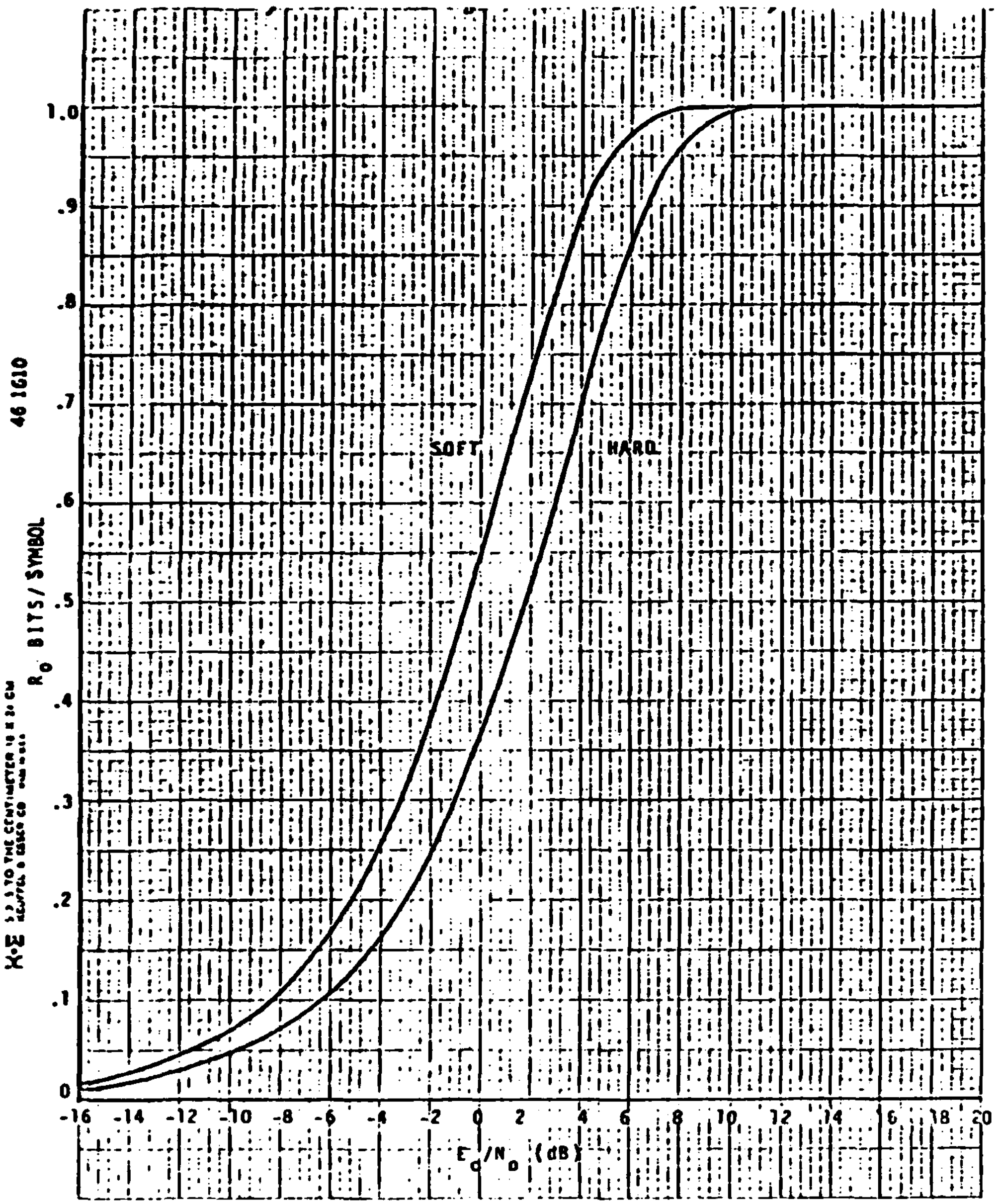


Figure 4.

Figure 5a $K=7$, $r=\frac{1}{2}$ Binary Convolutional Code - BPSK

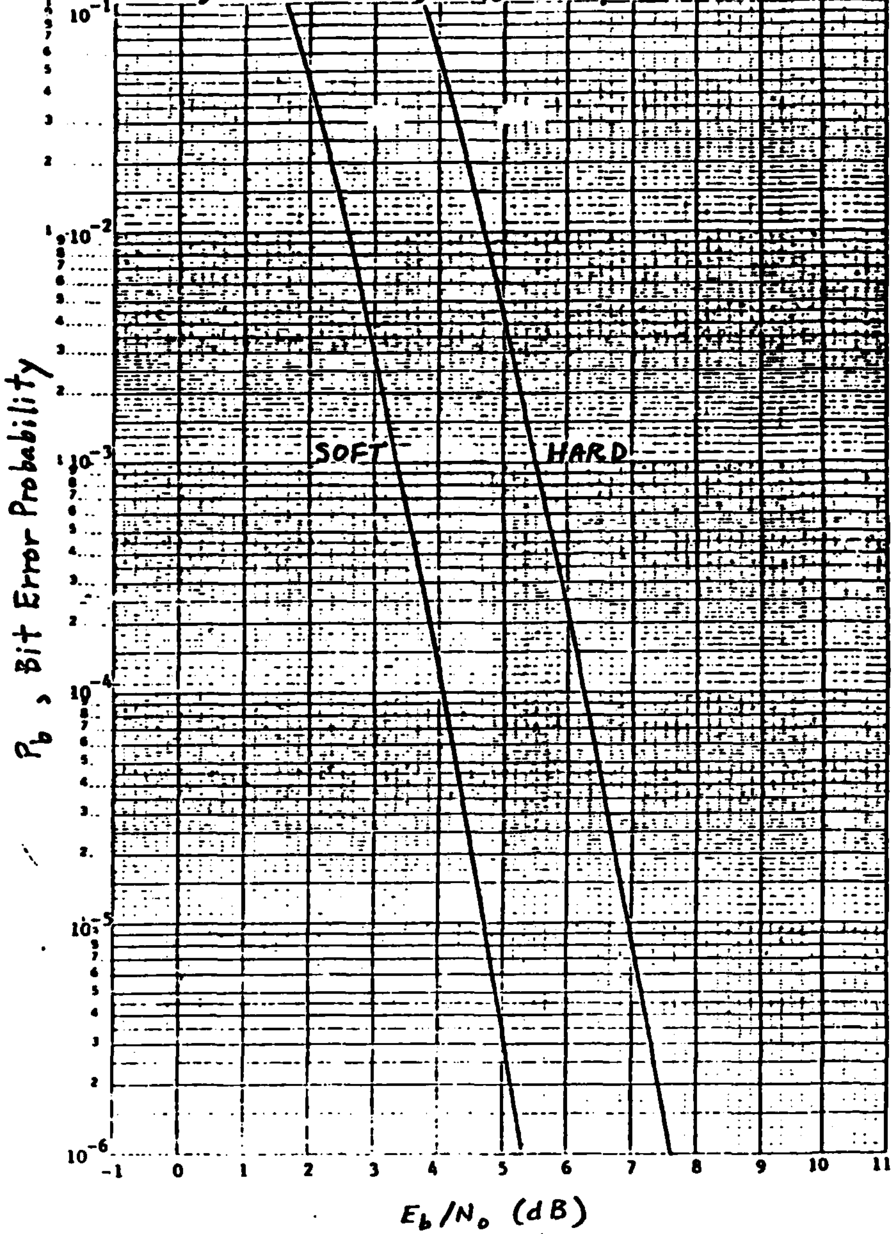
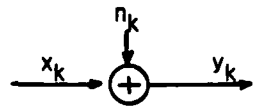


Figure 3.

Coded BPSK Modulation

$0 \rightarrow \sqrt{E}$

$1 \rightarrow -\sqrt{E}$



Repeat m Code:

$$0\ 0\ 0 \ldots 0 \rightarrow \underline{x}_1 = (\sqrt{E}, \sqrt{E}, \ldots, \sqrt{E})$$
$$1\ 1\ 1 \ldots 1 \rightarrow \underline{x}_2 = (-\sqrt{E}, -\sqrt{E}, \ldots, -\sqrt{E})$$

$$(E_b = mE)$$

ML Receiver:

- $n_k \sim N(0, \sigma^2)$

$$m(\underline{y}, \underline{x}) = y_1x_1 + \underbrace{y_2x_2}_{m(y,x) = yx} + \ldots + y_mx_m$$

- $n_k \sim N(0, \sigma_k^2)$

$$m(\underline{y}, \underline{x}) = \frac{y_1x_1}{\sigma_1^2} + \frac{y_2x_2}{\sigma_2^2} + \ldots + \frac{y_mx_m}{\sigma_m^2}$$

Figure 5.

Pulse Jamming

Suppose the jammer has average power J but can now concentrate more power in a pulse while still maintaining average power J. Then the equivalent noise density is

$$N_0^1 = \begin{cases} \frac{J}{\rho W} = \frac{N_0}{\rho}, & \rho \text{ fraction of the time} \\ 0, & 1-\rho \text{ fraction of the time.} \end{cases}$$

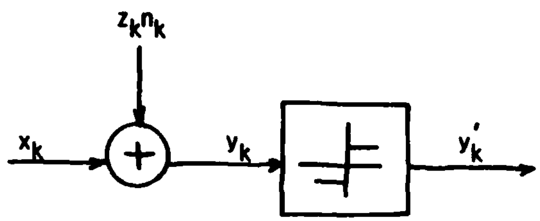
Then

$$P_b = \rho Q\left(\sqrt{\frac{2E_b}{N_0}\rho}\right)$$
$$\leq \rho \tfrac{1}{2} e^{-\rho \frac{E_b}{N_0}}$$

Figure 6

$\sqrt{E_3}$

$-\sqrt{E_3}$



$n_k \sim N(0, \frac{N_0}{2\rho})$

$Pr\{z_k=1\} = \rho$

$Pr\{z_k=0\} = 1-\rho$

z_k and n_k are independent

Unknown Jammer State:

$m(y_k, x_k) = y_k x_k$ or $y_k' x_k$

Known Jammer State:

$m(y_k, x_k | z_k) = x(z_k) y_k x_k$ or $c(z_k) y_k' x_k$

Figure 8

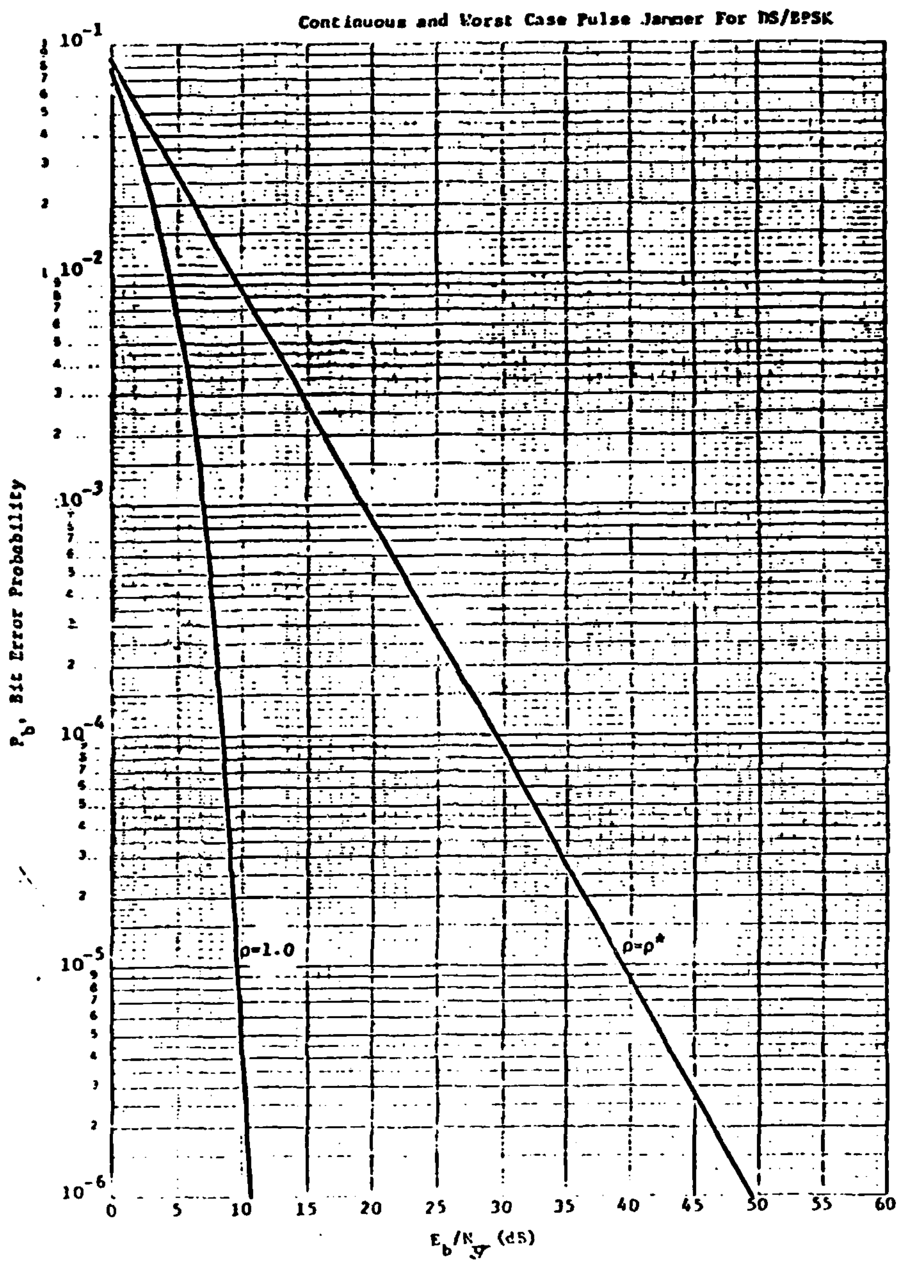


Figure 7

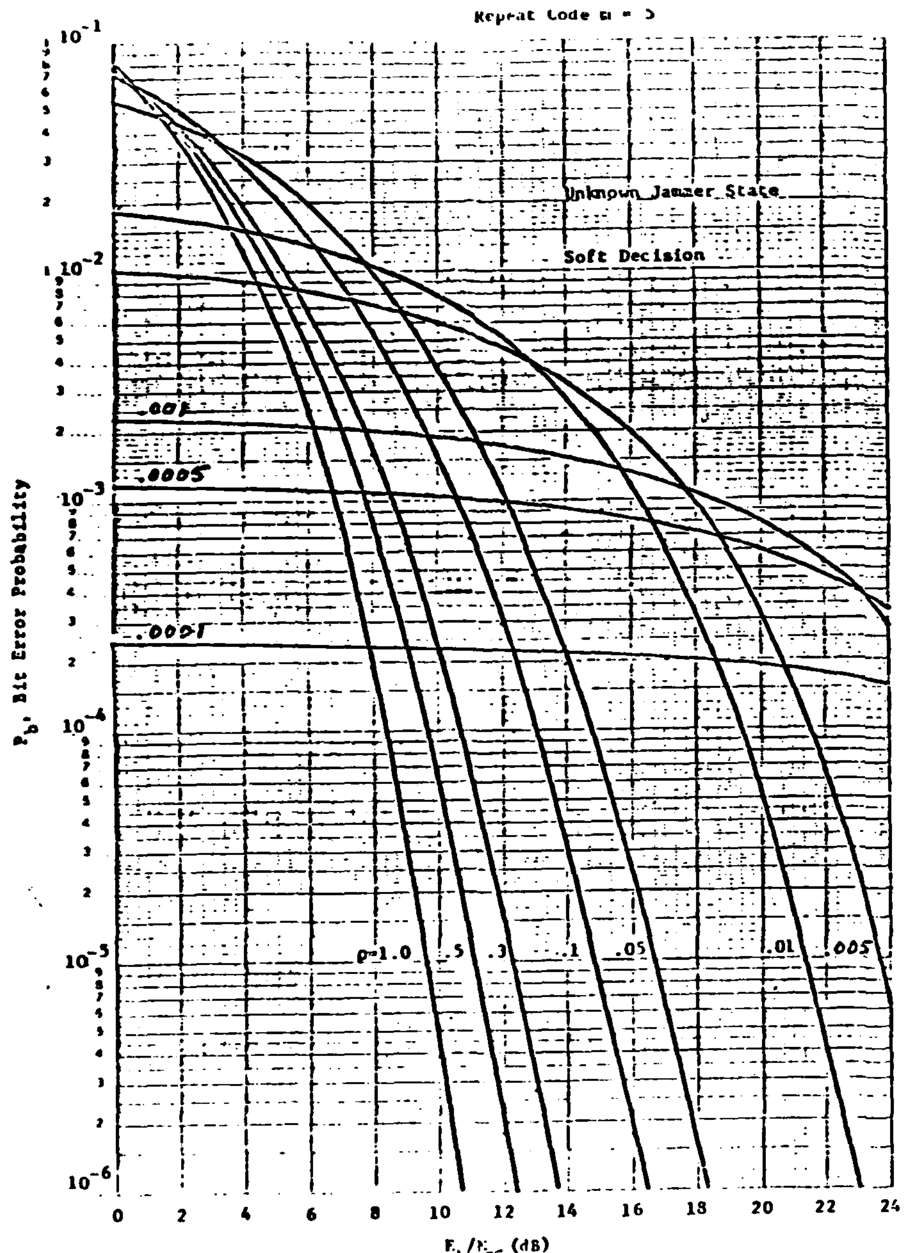


Figure 10

Soft Decision / No JSI

$$m(\underline{y},\underline{x}) = \sum_{k=1}^{m} y_k x_k$$

$$P_b = \sum_{k=0}^{m} \binom{m}{k} \rho^k (1-\rho)^{m-k} Q\left(\sqrt{\frac{2mE_b\rho}{KN_0}}\right)$$

Hard Decision / No JSI

$$m(\underline{y}', \underline{x}) = \sum_{k=1}^{m} y_k' x_k$$

$$\epsilon = \rho Q\left(\sqrt{\frac{2E_b}{mN_0}\rho}\right), \quad P_b = \sum_{k=\frac{m+1}{2}}^{m} \binom{m}{k} \epsilon^k (1-\epsilon)^{m-k}$$

Soft Decision / JSI

$$m(\underline{y},\underline{x}|\underline{z}) = \sum_{k=1}^{m} c(z_k) y_k x_k \qquad c(0) \gg c(1)$$

$$P_b = \rho^m Q\left(\sqrt{\frac{2E_b}{mN_0}\rho}\right)$$

Hard Decision / JSI

$$m(\underline{y}',\underline{x}|\underline{z}) = \sum_{k=1}^{m} c(z_k) y_k' x_k \qquad c(0) \gg c(1)$$

$$\epsilon = Q\left(\sqrt{\frac{2E_b}{mN_0}\rho}\right), \quad P_b = \rho^m \sum_{k=\frac{m+1}{2}}^{m} \binom{m}{k} \epsilon^k (1-\epsilon)^{m-k}$$

Note JSI = Jammer State Information

Figure 9

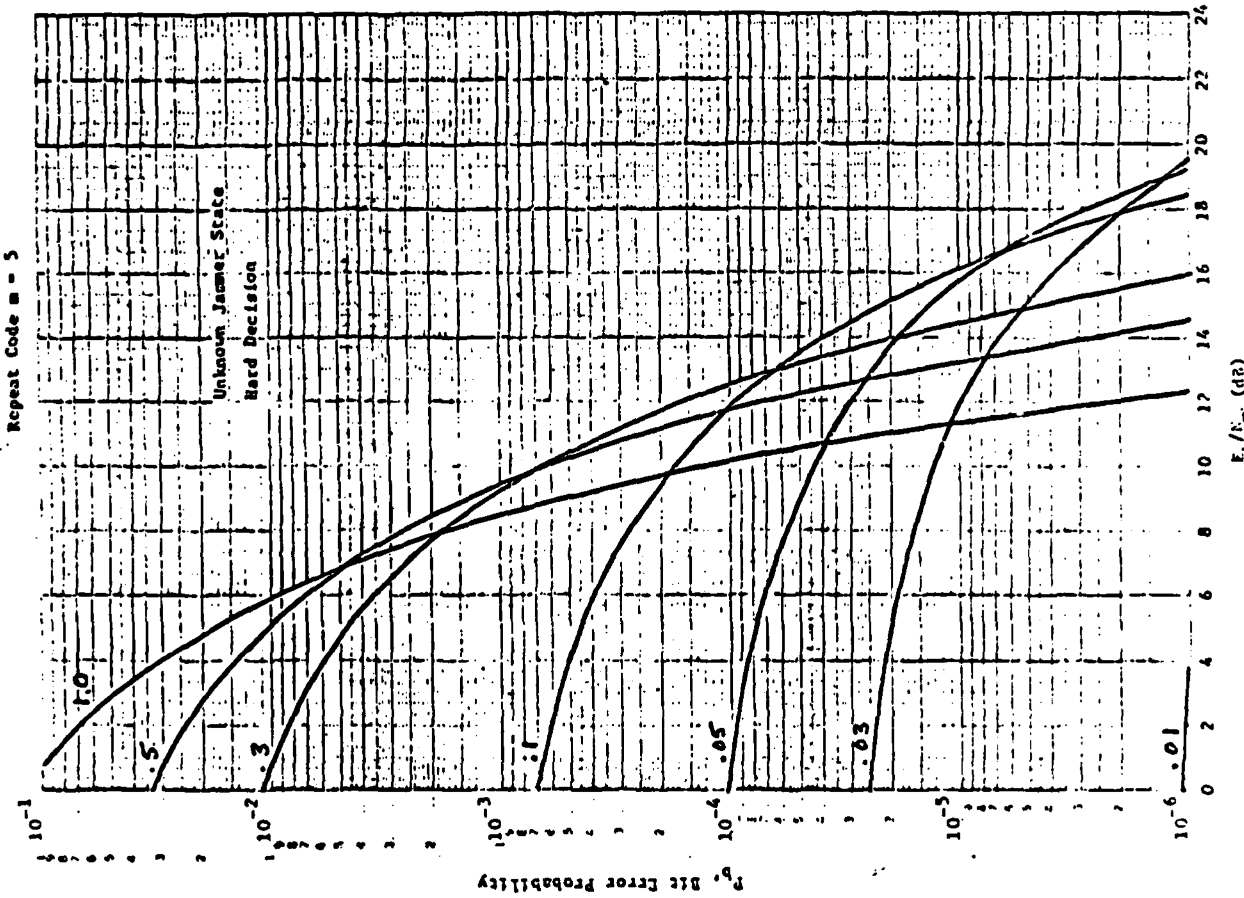


Figure 11

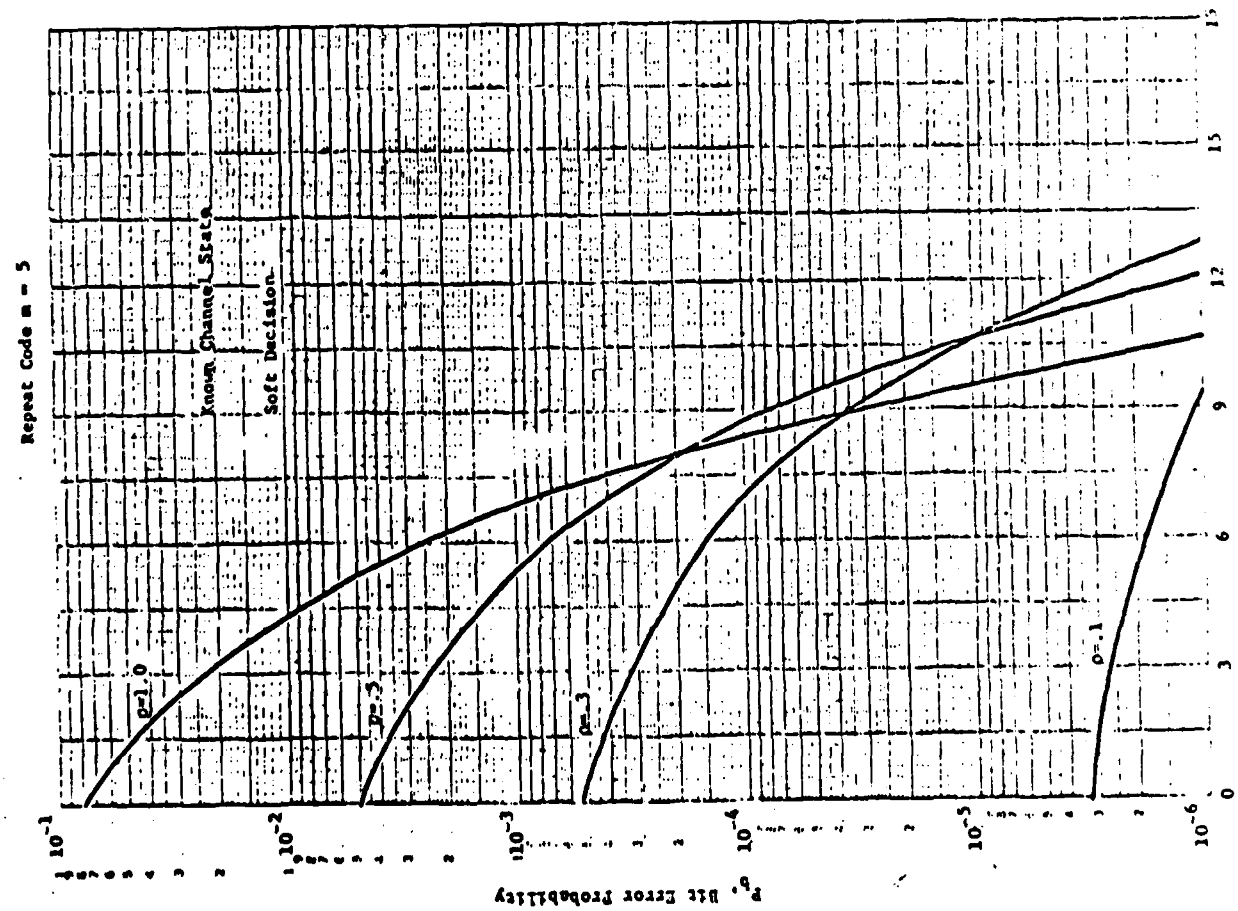


Figure 12.

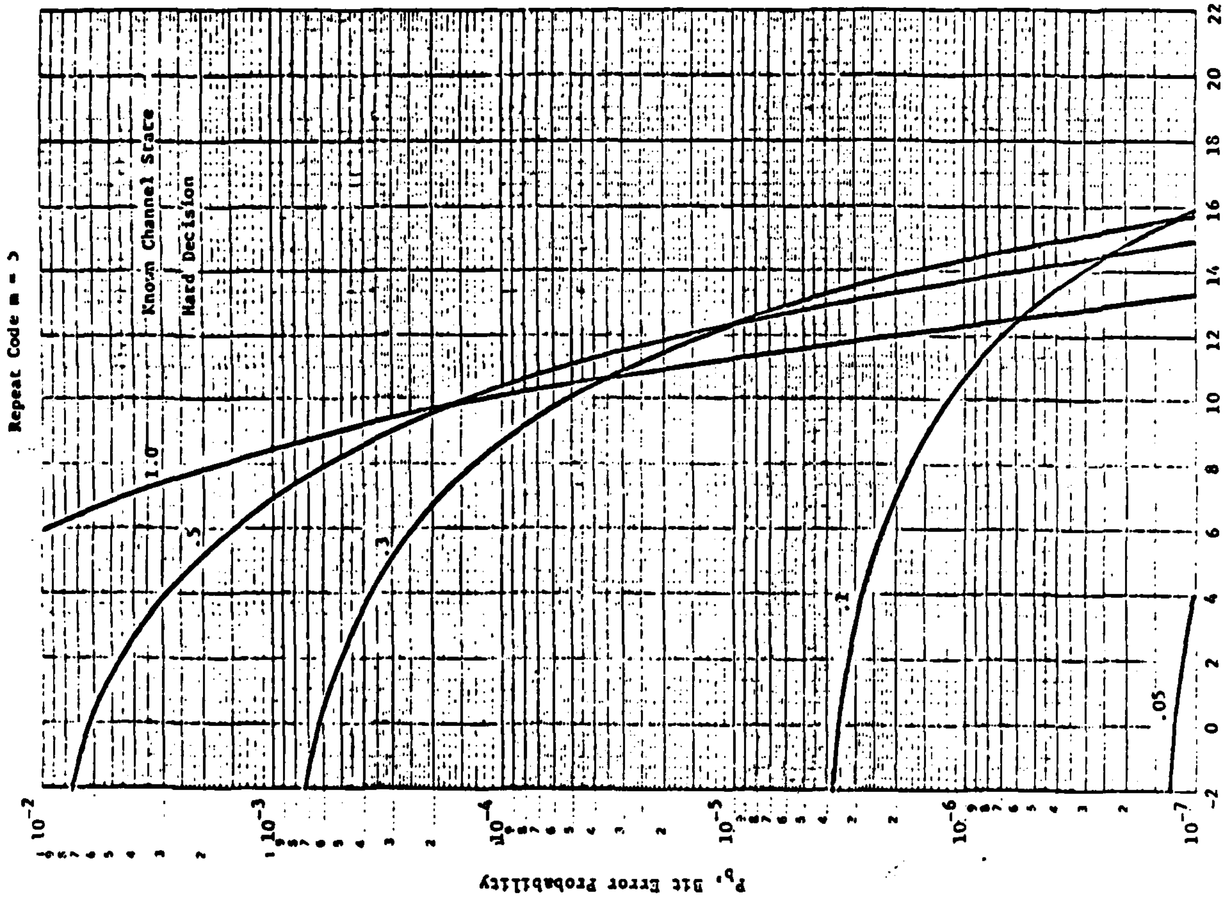


Figure 13.

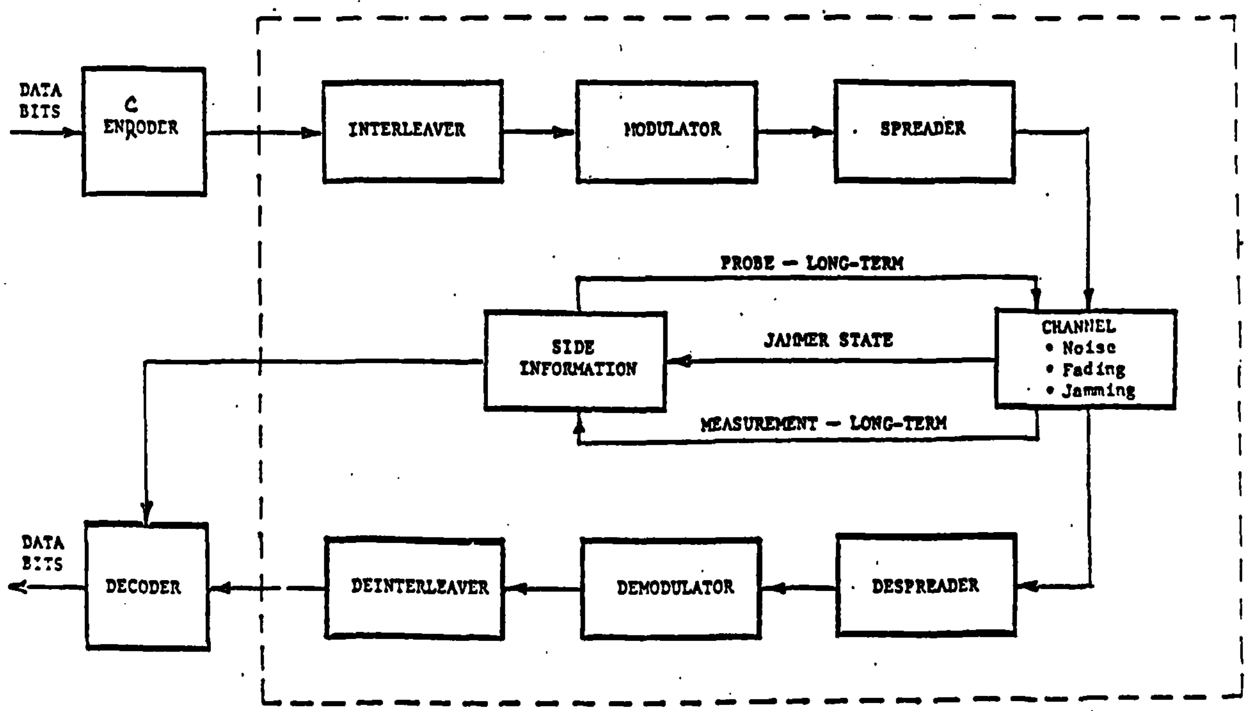


Figure 3-1. AJ System Overview

Figure 14

$P_b \leqslant G(R_0)$

$G(\cdot)$ = code-related function

R_0 = function of E_c/N_0

$E_b = rE_c$

Figure 15.

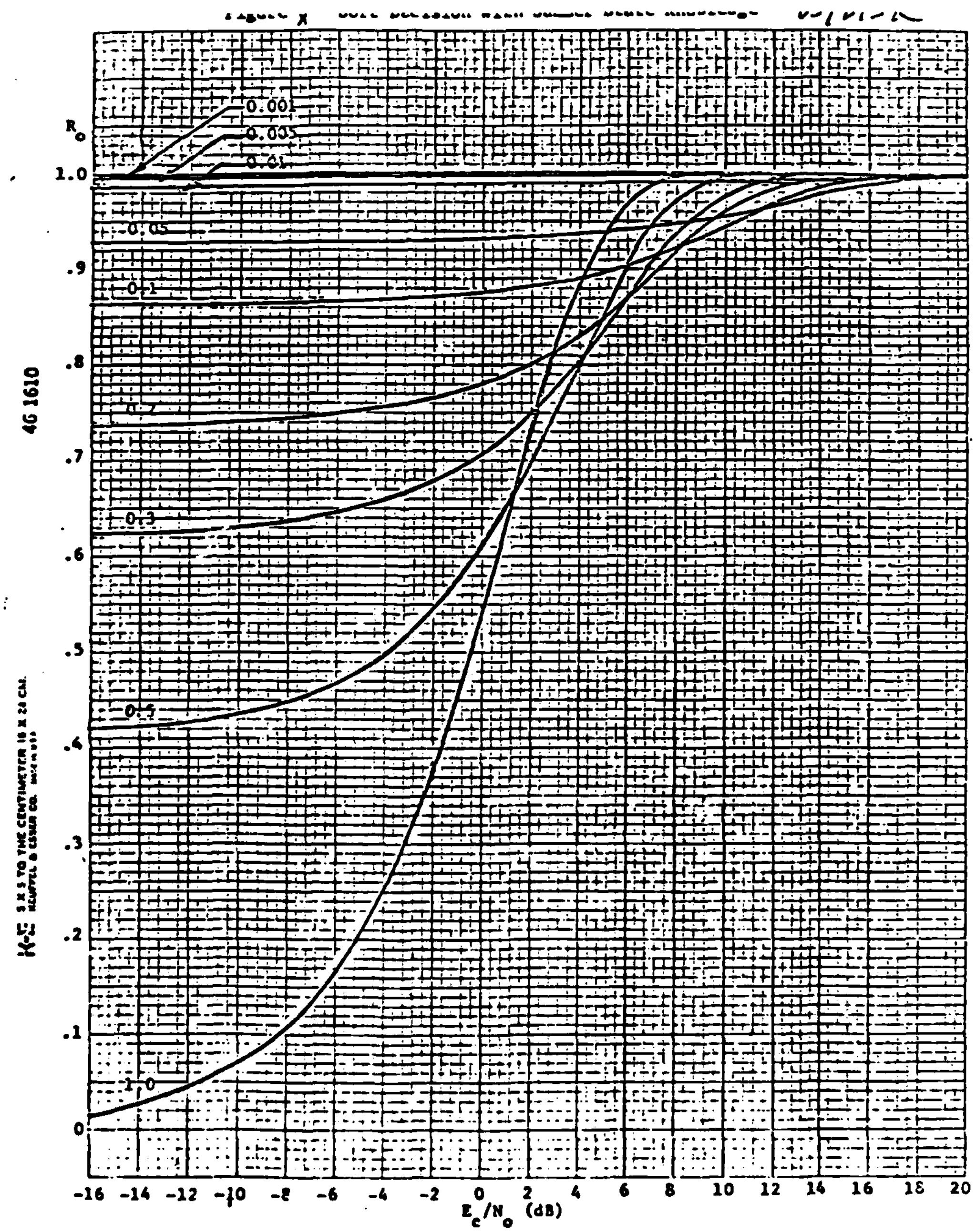


Figure 16

Figure 11 Soft Decision With No Jammer State Knowledge – DS/BPSK

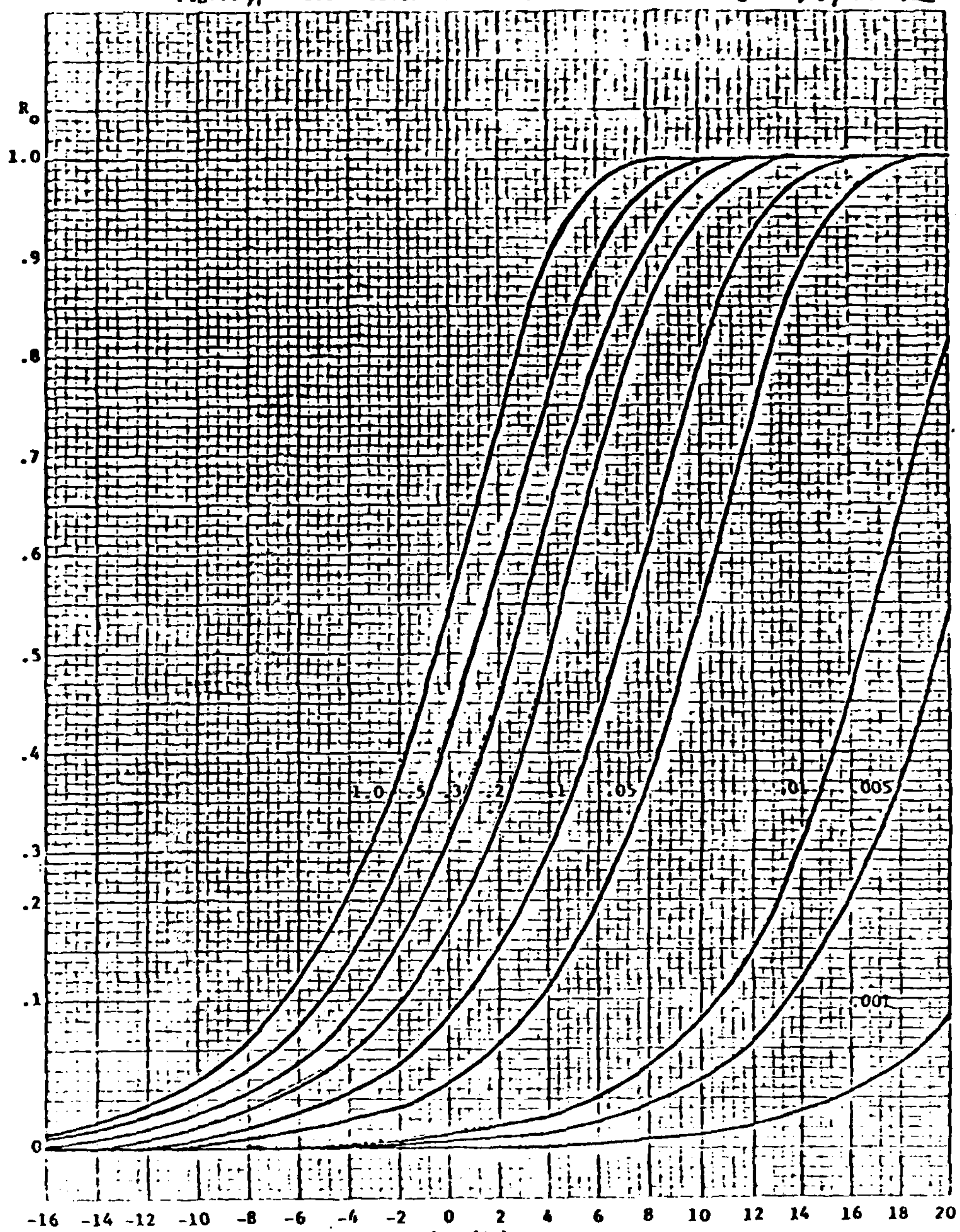
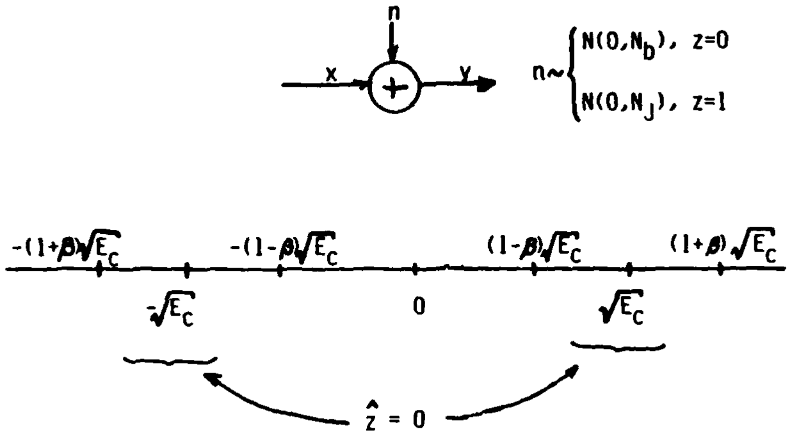


Figure 17.

Performance Analysis of Coded D/S
Spread Spectrum A/J Receivers with
Jammer State Estimates*
by

F. El-Wallly and D. Costello, Jr.
Department of Electrical Engineering
Illinois Institute of Technology



$$n \sim \begin{cases} N(0,N_b), & z=0 \\ N(0,N_J), & z=1 \end{cases}$$

$$N_J = N_0/\rho + N_b, \quad m(y,x) = \begin{cases} \frac{yx}{N_b}, & \hat{z} = 0 \\ \frac{yx}{N_J}, & \hat{z} = 1 \end{cases}$$

* GLOBECOM '82

Figure 18

A Nonparametric Mitigation Technique
For Partial Band Jamming of MFSK
Frequency Hopped Communication Links*

by

A.J. Viterbi
M/A-COM LINKABIT, Inc.

$\epsilon_1 \qquad \epsilon_{m^*} = \max_m \epsilon_m$

$\epsilon_2 \quad \epsilon_{m^*} > A \max_{m \neq m^*} \epsilon_m \implies \text{good} \quad (\hat{z} = 0)$

$\vdots$

$\epsilon_M \quad \epsilon_{m^*} \leq A \max_{m \neq m^*} \epsilon_m \implies \text{bad} \quad (\hat{z} = 1)$

$A \geq 1$

M inputs 2M outputs

Worst Case Partial Band Jamming

Figure 20.

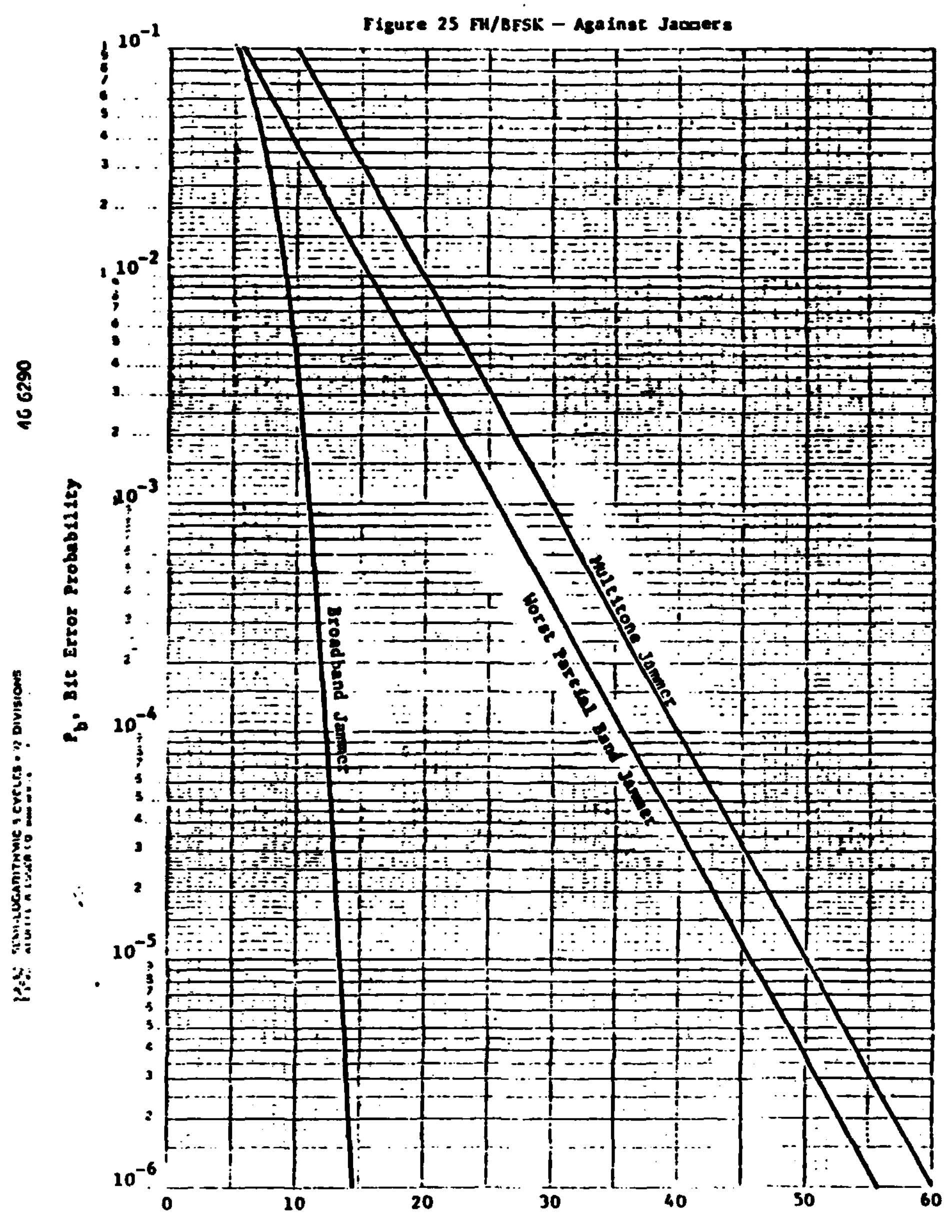


Figure 19.

* MILCOM '82

Performance of FH/MFSK with List Metric Demodulation Against Noise and Tone Jamming

by

M. Creighton
Ph.D. Candidate
UCLA

$$\begin{matrix} \epsilon_1 \\ \epsilon_2 \\ \vdots \\ \epsilon_M \end{matrix} \Longrightarrow \begin{matrix} \ell_1 \\ \ell_2 \\ \vdots \\ \ell_M \end{matrix} \qquad m(\underline{\ell}, m) = N_{\ell_m} \quad m=1,2,\ldots,M$$

ϵ_m $[0,\infty)$ $\ell_m \in \{1,2,\ldots,M\}$

Optimum Metric: $N_{\ell_m} = \ell n\, q_{\ell_m}$ $\ell_m = 1,2,\ldots,M$

$q_k = \Pr\{\ell_m = k \mid m\}$ $k=1,2,\ldots,M$

Partial Band Jammer:

$\Pr(z=1) = \rho \Longrightarrow q_k(1) = \Pr\{\ell_m = k \mid m, z=1\}$

$\Pr(z=0) = 1-\rho \Longrightarrow q_k(0) = \Pr\{\ell_m = k \mid m, z=0\}$

$= \delta_{k1}$

$\bar{q}_k = \rho q_k(1) + (1-\rho)\, q_k(0)$

Figure 21.

No JSI

$m(\underline{\ell}, m) = N_{\ell_m}$ $m=1,2,\ldots,M$

Optimum Metric: $N_{\ell_m} = \ell n\, \bar{q}_{\ell_m}$

$m=1,2,\ldots,M$

JSI:

$$m(\underline{\ell}, m \mid z) = \begin{cases} N_{\ell_m}(1), & z=1 \\ N_{\ell_m}(0), & z=0 \end{cases}$$

Optimum Metric: $N_{\ell_m}(1) = \ell n\, q_{\ell_m}(1)$

$N_{\ell_m}(0) = \ell n\, q_{\ell_m}(0)$

$m=1,2,\ldots,M$

Example: $M = 8$

$\underline{N}_1 = (10, 0, 0, 0, 0, 0, 0, 0)$

$\underline{N}_2 = (10, 8, 4, 0, 0, 0, 0, 0)$

$\underline{N}^* = (\ell n q_1, \ell n q_2, \ldots\ldots\ldots, \ell n q_8)$

Figure 22.

APPLICATIONS OF CODING TO SPREAD SPECTRUM COMMUNICATIONS

DISCUSSION

QUESTION: Is it true that sufficiently low rate codes will always force the jammer to $\rho=1$, namely continuous jamming or full band jamming. If so isn't this overall a good coding strategy?

ANSWER:(OMURA) Yes, that's generally true. The worst-case value of ρ generally is inversely proportional to the SNR in the channel. But that's the coded symbol SNR. So as you go to lower and lower code rates, the coded symbol SNR gets smaller and smaller for the same bit energy to noise ratio. And therefore you soon reach $\rho=1$ and the jammer can't exceed $\rho=1$. So generally that's true as you go to lower rate codes. Since you're getting more diversity, it's just a question of diversity. The more diversity you throw in, the more the effectiveness of the pulse jammer is weakened, and soon all it can do is broadband jamming or continuous jamming in the direct sequence case.

QUESTION: Please comment on the viability of feedback threshold decoding with interleaving in a pulse jamming environment for high data rates greater that 100 megabits per second.

ANSWER (ODENWALDER): Certainly threshold decoding is a simpler decoding approach, which if you're going to such a high data rate, you might consider. It's not as powerful as a soft decision approach but certainly when you go to such high data rates it should be considered. The other thing is the interleaving becomes very simple. There's a simple way of including it right in with the encoding. Basically you replace each stage in the encoder and the decoder with a register of some length that determines the interleaving level, so there's an easy way of implementing the interleaving. This makes an encoder/decoder look like a group of parallel encoder/decoders. So in that respect the interleaving comes very nice with the feedback decoding. I don't know about 100 megabits per second. That's really getting up there and you would probably have to have several decoders in parallel, even with something as simple as feedback decoders.

BERLEKAMP: We have 120 megabit machines in the field and they have no parallelism. It's an RS decoder, length 255, redundancy 12, with 243 information symbols. They were shipped in November of 1982

ODENWALDER: Feedback decoders are very simple. They are not like Viterbi decoders or sequential decoders by any means. Each are just a few chips. So even putting a few in parallel is still a very simple matter. Simpler perhaps than a single more complex decoder for another type of code

QUESTION: What are the ways in which coding/interleaving differ from matched filtering in spread spectrum environment?

OMURA: Let's just look at a direct sequence spread binary PSK system, or a multiple access system where each user has a different binary sequence on top of which he puts the binary information. The receiver can have a bank of matched filters that are matched to each of the sequences, or it could correlate at the receiver. Now, when we say interleaving here, we really mean interleaving the coded data bits which consist of many chips of the direct sequence. So the items discussed here are different in the sense that interleaving here is interleaving decoded symbols which consist of many PN sequence chips. I hope that answers the question. And you do need to do interleaving. You may have, for instance,

fading or pulse jamming that may last for the duration of several coded bits, which you would want to scramble up on a bit rather than on a chip basis.

RISTENBATT: I'm the one who asked that question. I'm thinking about such things as coherent detection. For example, if you are hopping over different frequencies, then you probably cannot expect to use coherent sampling. In other words, in the spread spectrum community now we're getting these CCD and SAW matched filters that are supposed to be very powerful, tremendous correlators. I'm searching for thoughts on the impact of these devices coming down the path, which will be our digital correlators (even which can do full-matched filtering). I'm searching for some of the rules or insights of when you would expect to use these devices, and when and how the two go together. It's a pretty open-ended question.

OMURA. My impression is that you use these SAW filters or matched filters of this kind when you're talking about codes of reasonable short lengths, which means that you're only using them in a multiple access environment, not in an anti-jamming environment for the most part. But maybe that's not true.

HUTH: First of all, we talked about jamming yesterday, and we talked about somewhat on-board processing and so on. There are implementations that use SAW filters in their demodulation process, and it is strictly a jamming environment.

OMURA: That's for primarily multiple access. On top of that is frequency hopping, isn't it?

HUTH: They are used strictly in a demodulation process, so it is not true. Their application has nothing to do with multiple access. It is for a frequency hopping processor. You could use it for either, but you can't say that they are not used in A-J environment because they are.

ODENWALDER: Typically you have to combine several of these chips in a frequency hopping system, to get the diversity.

WOZENCRAFT: If I could provide a slightly different answer to that, my own opinion would be that you would use the SAW device and the direct spreading sequence to give a long enough pulse duration so that the difference between the coherent performance and an incoherent frequency-hopped performance would become negligible. So I disagree with Dr. Jain's conclusions of yesterday. I think that you very definitely do want to use both frequency hopping and direct spreading. And I think they fit together very nicely.

SIMON: Let me raise another issue which I brought up in Florida in the context of workshop there but I think has interest here. Jim and Joe were talking quite a bit about optimum metrics, finding robust metrics for jam channels. Most of the attention has been focused on the metric, or if you like, the decoder. And very little attention has been focused on the encoder. Now we have been doing a lot of work for ISI channels (which are memory channels), on what happens if you do a search for optimum codes and demodulator-decoder combinations for such memory type channels. We found that indeed, on channels with memory, there's a lot to be gained by looking for codes which are optimum in that kind for a channel (as opposed to the well-known codes that Joe found for the additive white Gaussian noise channel) or interference. Or partial response channels. (We've looked at those too.) I think there's an interesting area here to look at, and that is on channels that are non-stationary, non-Gaussian, like the kinds of channels we are talking about. I think some attention should be paid to

searching for optimum codes and of course optimum decoders too, on such channels. I have an interest in this problem. I'm just trying to raise it as an issue. I don't know how much there is to be gained, but I think that it is something to be looked at. Does someone want to make a comment?

LEINER: I'd be interested to hear your reaction/opinion about the applicability of that when you have spread spectrum which often allows you to resolve or measure the channel. I assume the intersymbol interference you're thinking about is the kind that's caused by multipath. When you have spread spectrum and cannot resolve multipath components, you may have intersymbol interference in that the components may overlap, but nevertheless you can still pull out each bit's contribution. Would the kinds of techniques you are looking at be relevant in that case?

HUTH: Let me make a comment. If I understand you Marv, you are not talking about ISI in this case. All he's talking about is, we've got an environment that is not additive white Gaussian, it has a lot of different things in it. Why are we using codes that were designed for AWGN, or why aren't we designing codes for this environment? Indeed, you have to get rid of the susceptibilities but there's been no real work done which says, hey, this is the type of environment, it's not Gaussian. Marvin's case was ISI, but in this case we're talking about ...

SIMON: It's really two separate issues. I'm just saying that in Florida I talked in context from an ISI channel I'm saying now let's talk about the Gaussian non-stationary channel, and again, the issue is what is the optimum code for that kind of channel.

WELCH: Well, it's really more than just the channel, because you're method of spreading also is going to control what sort of a channel the code is seeing.

Tail-wing? Ask STEIN

STEIN: Yes, but we've talked a little bit about nuclear effects, we've talked about a number of things that are not necessarily intelligent situations. It's not clear that there is a good code for the simulation cases either. There's not been work done there. I assume that many of you are familiar with the work that Dave Chase and some other people did under the tail-wing program and I'm really curious not to have heard anything about that approach here today, where the topic is coding. What Chase pointed out is if you have a TW space you have a TW product per bit. That's a 1/TW code rate, and we all have learned, there is something much better than sheer redundancy in the fading world where everybody traditionally used diversity and now recognizes it's much more beneficial to use a designed code. Chase proposed and did some exploration on efficient codes to fill the TW space and I believe he has continued to do some work under Air Force sponsorship although I'm not familiar with it. I am curious if anybody is. I'd love to know where that's heading, whether anybody has explored its limits.

WELCH: Would that have applications other than white Gaussian noise channels?

STEIN: Sure would. Just as coding applies to any burst channel.

HUTH: Wasn't that oriented towards the HF channel and channels like that though (and I'm not saying it doesn't have applicability to other situations), and you do have somewhat a different channel. But work was done in that direction by Dave. I don't know what he's doing now, I haven't seen anything come out of their organization.

OMURA: In general, for large TW

products, if you're talking about a code in the sense of spread spectrum being an error correcting coding scheme, you'll find that it's a very low rate code. Here you very quickly reach a point of diminishing return in dropping a code rate from a half to a third until pretty soon it doesn't make any difference. So why use such an extremely low rate code. Why not use something like some direct sequence spreading and then code on top of that with a convolutional code of rate a half or a third. That's close enough. You don't have to go to a rate 1/TW which is an extremely low rate code. Maybe I'm missing the point.

REIFFEN: Your argument is good enough. But what Marv is asking, what Seymour is talking about, what Chase is doing, is to say let's not just say that's good enough without making sure we've done a study.

OMURA: Another thing you want to confirm is that your system is robust enough. A system that's going to be designed well for one situation may not work well in others, and you've got to be careful.

REIFFEN: I'd like to make a comment primarily on the first question Jim Omura answered having to do with low rate codes and values of ρ. We saw the wide divergence between the performance of anti-jam systems for $\rho=1$ and ρ less than 1 when the probability of error that we're looking at is very low. In fact, it get's worse and worse as the probability of error gets lower. I think that all of this was implied in what was said. But I'd like to say it explicitly here. What that means is that when we apply our error-correcting code, we want to do it on a raw channel, which if we actually made a decision, (and I'm deferring the issue of hard decision versus soft decision) actually looks like a noisy channel. Because we noted that when the channel is noisy, the difference between the $\rho=1$ and the optimum ρ is not very much indeed. So we want to force the jammer to operate on the raw channel which is noisy, and then we take out that noise with the antenna and coding. I think that's approximately the general rule in the design of robust anti-jam systems. It ties together the intimate relationship between modulation/demodulation and coding. Good A-J systems have got to have coding together with appropriate choices of modulation or demodulation.

PURSLEY: I think that's a good point. I've done very limited studies but I think it would be advisable to do more. The studies I've done have to do with looking at Reed-Solomon codes in diversity, where you have a tradeoff between the rate of the code you're using and the diversity level. And surprisingly, there does seem to be an optimum combination for each block length code. It isn't down to the low rate Reed-Solomon code it's sort of a medium rate Reed-Solomon code with certain order of diversity to get the rate down. I think that should probably be done for other classes of codes as well. My interest has been limited to block codes, in particular Reed-Solomon. Maybe you can comment, Joe, on the things that might have been done on convolutional codes along the same lines.

ODENWALDER: The only experience I've had is with convolutional codes and I didn't notice that much of a difference in the few cases I've looked at. For example, the small performance improvement of a rate 1/8 code over that of a rate 1/4 convolutional code with 2 symbol repetition wouldn't justify the need for a slightly more complicated decoder. Plus there's another reason and that is, quite often you would like to be able to handle different data rates. You'd like to have, for example, the same hop rate but combine different numbers of chips to get different data rates for the disadvantaged and

advantaged users. Repetition is an easy way of doing it. If you do not use repetition you have to accommodate different code rates and it's much more complicated. The repetition idea probably does deserve more attention.

BERLEKAMP: We were involved in one aspect of the JTIDS program. I'm sure people here might know more about the overall system than I do. From what I saw there I got the impression. that the reason they had chosen this 32-ary alphabet as opposed to a bigger one like a 64-ary or 128-ary was really the SAW's limitations and not coding. Now the coding gets better if you go to bigger and bigger alphabets (at least from my point of view). *The design decisions were made 8 or 9* years ago, and the program is still bumbling along.

HUTH: The program is still doing the same thing but the decision for that size was made when they came up with the *program and I think if you started with a* program right now you'd have different numbers.

BERLEKAMP: I also think they greatly overestimated the coding complexity that they were looking at.

HUTH: We talked about binary codes and there seemed to be a lot of effort there. And I think Joe made a comment about the the dual K or triple K codes. I guess somewhere along the line I did not understand why M-ary codes have not been built or attest to an extent anywhere nearly as much as the binary. Especially, if I'm reading it right, I believe the list demodulator that Jim Omura is talking about is applicable to M-ary codes not binary codes. There's work being done in this area, but I don't see what happened to it.

ODENWALDER: There was very little done on dual K and triple K. At least one dual-K coding system was implemented. Recently things have swung the other way toward the binary codes because they are easier to implement and for a given performance and complexity you are, in general, better off with the binary.

BERLEKAMP: It's a question of your likes and your speeds and all kinds of other things. But you've got to get your non-binary symbol form, and the alphabet it has to be fairly big. You can't do much with soft decision information. These are all drawbacks. You have to have a pretty smart front-end, you need a SAW or something similar out there to get your orthogonal demodulation.

BERLEKAMP: Just like Jim's list demodulator. I think there are a lot of things which, when you start to talk in a jammed environment have some good points. If you put all that together I don't see why we are so pushy. Even in soft decision implementations I don't see why we are pushing binary so hard for these *kinds of channels, when the modulation* structure is M-ary anyway

OMURA: Recently, Massey's paper seemed to indicate in some cases that you should at least think about looking at the binary equivalent, instead of going to an *M-ary channel, and you can come up with* very good codes that way.

REIFFEN: A word about choice codes. The bottom line that choice of code relates to is the E_b/N_0 performance which you get at the stated end-to-end error rate that the system requires. The fact of the matter is that people know all kinds of codes that produce interestingly low E_b/N_0, and the quest for better codes is diddling with the last few tenths of the dB. I think what I would observe in the area of codes is that we are close to the point of diminishing returns and that's why maybe coding theorists are very excited about it, but system engineers are not.

PURSLEY. I disagree. Because by talking

E_b/N_0 you're already specifying some sort of a channel which might not be true channel at all. There are lots of other parameters about these channels, that ought to be optimized.

STEIN: I think I have to emphasize again that it's not the stationary flat-noise environment that you're worried about. If there's ever to be any real use of wide-band direct sequence codes in the lower bands where there's heavy and very dynamically changing interference, I would put my bet on coding rather than on adaptive interference cancellers as an inexpensive implementation. If you've looked at the problems of any kind of adaptive interference cancellation those things are complex. I think people still don't really understand how they work, despite all the algorithmic work. It get's to be a self defeating problem. Coding almost looks like an automatic way to do it if you can tolerate enough errors. I do believe that there'll be a minimum code bit error probability. I'd like to know what it is I have no idea where that is.

PRICE: I'd like to go back to Shannon's original concept of efficient communication Wlog (1 + S/N), and all that. I wonder if anyone remembers that he proposed noise-like signals for actually achieving that capacity. And I don't think he had jamming in mind, but we know that PN is a goody against jamming or has been. Thinking now in terms of frequency hopping, I think it has been quite popular to use MFSK on top of frequency hopping I really wonder whether that is such a good idea. Of course it's simple enough, but why don't we go to hybrid PN frequency hopping with the idea that the PN is a Shannon type PN, for getting E_b/N_0 if you like, and that's also a crypto-type PN so that it's not easily jammed. And as a particular example, I'm thinking of the Ungerboeck "multiphase" codes which are also modulations that have good Euclidean distance. They probably are very difficult to demodulate with soft decision but I wonder if that isn't the step in the direction of living in both the natural world and the jamming world. So I'm talking about different kinds of baseband modulations for frequency hopping in particular, and getting away from MFSK.

ODENWALDER: I think I'd be leary of some of these multiple phase type codes in fading environments like nuclear simulation. I'd try and stay away from them.

PRICE: I'm adhering to the Shannon model here.

HUTH: You say MFSK, you could do MCSK and use those kind of symbols and use an orthogonal signalling set rather than using different frequencies. Actually you use the codes themselves, and that's part of the JTIDS structure.

PRICE: Thank you. I might have mentioned the MFSK functions as orthogonal functions in that context. But I don't think they are so good A-J wise.

REIFFEN: Let me talk about waveforms as well as codes. When I previously referred to E_b/N_0, I really meant by N_0 the ratio of J divided by bandwidth in a spread spectrum system. Because spread spectrum systems work against the jamming rather than the noise. In that context of E_b/N_0 we want to pick modulation and coding combinations that produce interesting low E_b/N_0 and we've already seen that simple inelegant, non-Shannon-like MFSK achieves interestingly low E_b/N_0s in which case the quest for better designs really has to be motivated by the promise of lower E_b/N_0s, and I hold that out as a hard-nosed challenge to anybody who comes up with ultimate codes, or ultimate modulations.

OMURA: I agree with that. I think you could come up with some better codes

perhaps for M-ary alphabets but I doubt if there is going to be much reduction in E_b/N_0.

WOZENCRAFT: Just to add another comment along those lines. it seems to me that where the search for good comes in. is much more into robustness in the face of variations in operating environment, in terms of specific E_b/N_0 however defined against a particular threat.

LEINER: A general comment. This discussion sounds a little bit funny to me. It sounds as if everyone is advocating one particular solution that's going to solve all the world's problems. As Marv was saying, you really have to take a look at what the right thing to do for the right environment and when the engineer goes out there to build a system he's got a certain set of user requirements to meet. So for example, if you're trying to tackle CW *interference* it very well will be that an adaptive interference canceller, or some some sort of cancellation technique is right. If you're trying to handle noise jamming, I guess it's wrong, except if you want to use an adaptive antenna, which is right for that situation. Although frequency hopping may help you with narrowband interference it may be wrong because you have to handle some LPI considerations for example, or because you are worried you may not want to use FSK for example because of some delay. If you put all the requirements on the system at once, you're going to wind up with nothing. So a discussion that says, this system is better than that system, I think is missing the point.

STEIN: Right now we are in a symposium where we are talking about research directions. When I think of research direction, I think of something that may have application long after I'm gone. Right now I think there's more known than we know how to implement. And certainly with regard to the design of systems for the 1990s and in the face of what we know about the threats for the 1990s, if anybody asks me, I say, "In the lower bands design a frequency hopper, direct sequence makes no sense". When I try to look beyond and I think about what has come to me, at what seems like the ultimate problem, of detectability and targeting, I see no alternative but to use the full bandwidth the whole time. I don't know how to do it. That's really the direction I guess my comments have been pushing towards. That's where research should be. What can we do to make that viable, because I think it has to be the ultimate solution assuming we are still fighting the cold war.

OMURA: Just one comment about the list metric that I mentioned at the end of my talk. Part of our motivation for that was to look at an environment where you had multiple users perhaps with different signal power levels, coming into a single receiver where you *simultaneously* demodulate, decoding many users at once. That kind of highly crowded environment where you really want to pull out many users in that same bandwidth, that was the motivation for looking at the list metric. Rather than other codes, one might want to look at how one might use the information about the environment I am hopeful that this might lead to some improvement.

WELCH: Could I ask a question of Joe? You were talking about the tail-biting conversion of a convolutional code into a block code in the decoding process. Were you suggesting that you double the received data, you make two copies of it and decode the double length thing and then take a look at the better selection? What happens if when you decode it isn't periodic. So when you get back to the same point in the data you get a different message bit?

ODENWALDER: The initial and final states are guaranteed to be the same. The paths in the middle are not necessarily merged of course, but that doesn't mean you have an error. Occasionally you will have errors. There's no way around it.

HUTH: One of the questions I have if you use the tail-biting argument like that is: Why are you using Viterbi rather than say a maximum likelihood bit decision, two-way recursion? It seemed to me that would be more appropriate for what you're doing.

ODENWALDER: Yes, you might consider doing that. But there might be basically continuous data that you have to decode, as well as some other short data. I'm thinking of some military systems where there are short messages as well as data that's essentially there forever. If a Viterbi decoder is used for the continuous data, the short messages are most easily accommodated using a similar path maximum likelihood approach.

OMURA: Also I don't think you buy that much by going to a complicated minimum bit error decoder, so why bother with it? I don't think it buys much in E_b/N_0.

HUTH: There are other things besides E_b/N_0 as we've discussed. There are some advantages to using that code but it depends on the environment. I was just questioning why you had gone to Viterbi decoding when you could have gone to the recursions. Joe showed one of the applications where you could do both recursions. Most of the time you're not in a position where you can block the data like that.

ODENWALDER: Actually in the decoding algorithm there are a lot of variations you could use. I said you could decode two or more times and that's of course a speed disadvantage. Your decoder has to operate faster. If you're clever there are other things you could do. You can look at the ending metric at the end of the first path and see which one is largest, and trace back. And there are all sorts of variations that you could use to avoid that speed factor increase. They use the same basic idea.

POSNER. I'd like to ask a question of Joe. Did you find the Golay code the first rate 1/2 code you tried or did you fiddle around?

ODENWALDER: It's been brought to my attention that Gus Solomon[1,2], has looked at some of these and there may be some of these around. It's my opinion that there probably are some other that are equivalent to the Golay code.

POSNER: You found it the very first code?

ODENWALDER No. I only looked at the optimum convolutional codes with constraint lengths of 9 or less.

POSNER: And they weren't all Golay? Only one of them was?

ODENWALDER. That's the only one that's optimum in that sense. If you back off from the optimum convolutional code you may be able to find another tail-biting Golay code: There's a good chance that for non-optimum convolutional codes, you could find another one that's equivalent to the Golay code.

RISTENBATT: In the spirit of what Stein mentioned about the research directions, that tended to trigger me to ask, do you think there's any merit in new problems in codes, searching for codes to do the best job in connection with adaptive arrays in spread spectrum? Maybe this is a place that coding people should be looking at.

OMURA: I'd like to comment on that one. One of the reasons for looking at the list metric is the following. Suppose you were in an environment where you have

multiple signals in the same bandwidth, perhaps like an adaptive array antenna, you might be able to learn something about the channel conditions in terms of relative signal strengths. For instance, if you take the extreme theoretical case where you have two transmitters, but one is near and one is far, and both using the same bandwidth and the same MFSK modulation. Here just knowing the distances (one is going to give you strong pulse, and one is weak) you should be able to resolve them perfectly. You can probably resolve 3 or 4 if you have exact information about the location and the energy level, it is all deterministic. So the question comes up to add noise and random variation, how much of that additional information helps you actually in the overall decoding process in resolving the different users. Like the adaptive receiver in the telephone channels where you have adaptive equalization, perhaps one can start to develop a histogram, learn about the channel propagation signal levels, develop knowledge of that, and use that knowledge to resolve the different users in the same bandwidth. That was the motivation for this metric also.

KRASNER: I have an overall comment about this business about coding with lower rate codes and so on. If you take the frequency hopping environment, for one thing, people are talking about memoryless channels. I think that's a bad assumption. You consider the fact that you have different types of interference and jamming if it's a heavy environment, some of your interferences are going to be stationary and some of them are going to be jumping around perhaps. Ones that are stationary, may not be jamming signals, maybe they are. Ones that are jumping around might be. What you would like to probably do is have some kind of process to adapt your coding in some way to basically not transmit data at those frequencies that represent constant interference. You might be able to model the remaining channels as memoryless because it's perhaps a jammer that's jumping around. There seems to be a lot of room for research in those areas.

OMURA: A lot depends on how fast the jammer is moving around and how fast you can learn about the channel conditions. Certainly for natural phenomenon like the HF channel, you send a sounder across the band and you know which channels propagate well, and which ones don't and of course you would transmit (hop) only in those bands that propagate well. I had a Ph.D. student, in fact, study that for the HF channel and then do a min-max problem to determine what's the worst kind of jammer in that environment.

KRASNER: Of course the jammer would play a two-way game. The jammer could be stationary in some places and jump around in others.

WELCH: Also somebody was mentioning yesterday that there was a philosophy that it was bad to react to the jammer. That is, you set up your system so that whether the jammer is there or not, you don't want to let him know that he's being successful.

LEINER: One clarification to that. It's only bad to react to the jammer in a visible way.

DUPREE: I have a very simple-minded question about adaptive arrays and interleaving and so forth. My simple-minded assumptions are these, that since we are processor throughput limited in adaptation time in the present state of the art for adaptive arrays on satellites, that we have to rely on interleaving to cover the gap between the time the jammer comes on and the time that we are fully adaptive or we can suppress the jammer. We have made the assumption that (a) the interleaver will fully take care of this time interval, i.e. if we say hypothethically 1/10

second of interleaving time and convergence time is required, that the interleaver fully bridges that gap and there's no loss or disruption in the communication. (b) The next thing we get to is, perhaps we can't quite meet that design requirement and so the program manager says," Oh well, it doesn't matter. You're going to lose some of your communication." My question is: "Is the assumption (a) correct, and (b) is there a disruption beyond the interleaving period, that is, is there going to be propagating effect in the decoder as a result of perhaps not adapting within interleaving time?

ODENWALDER: I guess if you have a long burst that get's through you certainly could. The decoder could lose sync, if the decoder actually does synchronization on its own. For example, the decoder would have to know node-sync. If it does that on its own and you give it a long burst it could throw it out of sync.

WELCH: If anyone else has any other question, we do have a free-for-all tomorrow morning, at which time you can ask them.

REFERENCES

1. G. Solomon and H.C.A. van Tilborg, "A Connection between Block and Convolutional Codes," SIAM J. Appl. Math., Vol. 37, No. 2, pp. 358-369, October 1979.

2. Richard W.D. Booth, Mark A. Herro, and Gustave Solomon, "Convolutional Coding Techniques for Certain Quadratic Residue Codes," ITC Proceedings, pp. 168-177, 1975.

SESSION 4 - SPREAD SPECTRUM NETWORKS

BARRY LEINER:

The session this afternoon is spread spectrum networks. We've heard a lot over the last couple of days about the various different things you can do on a communication link that use spread spectrum, ranging from antennas and coding, signal processing, etc. What we are going to be talking about this afternoon is how spread spectrum can be used in a network and what the impact of using spread spectrum in networks is on both the requirements for spread spectrum and the requirements for networking. In particular the question I put to the panel is: In what way is a spread spectrum network different than the simple addition of spread spectrum and network ? What synergistic properties happen that either add or subtract from the capabilities you might expect to get when you have spread spectrum separately and a network separately ? The panelists and the talks you are about to hear are listed here. Figure 1 I guess what I'll do is just launch into my talk and then I will introduce each speaker as they come on.

What I'm going to talk about is layering in a spread spectrum network. The idea of a layered architecture at first blush leads you to believe that you can simply slip spread spectrum in at the appropriate level, namely the link and physical levels, and basically not affect the network level and above. That's a nice naive notion. I actually hope to convince you that is partially true. As opposed to the ISO-layered architecture, this is what you might call the Leiner layered architecture. Figure 2 It's a six layered model and the notion is the following. If you view things in a modular way, data flows from an application to an application by using services of a transport level which provides bit transport; reliable end-to-end communication between the two end points requiring communications. They make use in turn of an internet layer which allows different networks to be glued together. This layer is dotted. That's meant to indicate that there may be other internet entities in here such as gateways between networks. The next level down is the network layer. Each node in a network has to implement some sort of network layer protocol which has to do with routing and congestion, and basically how you move the data through the various nodes in a network. The network layer in turn makes use of link layer protocols which establish what may or may not be reliable but at least is communications between the various neighboring nodes in an network, nodes that can communicate directly by the physical media which is the bottom layer of the architecture. The way data flows through the network is, it starts *out at the application level and wends its* way down through the various levels of the originating entity which in this case you might think of as a combination of host computer and originating communication device. As it proceeds through the network the data works its way up to the network level and then back down as it decides which link it needs to go out, the next node etc. until finally you wind up with the end device where it gets delivered to the application. Notice at these levels there's end-to-end communication between the protocols. This is a nice theoretical model. By the way, Barry Levitt made mention that he did his curves on his IBM PC. I did this viewgraph on my TOPS20 that sits effectively on my desk via networking.

Let me give an example. Figure 3 A typical application that you may want to use communications for is electronic mail. You want to transport it using an end-to-end reliable communication protocol, in a

typical case that would be TCP (transmission control protocol). The internet protocol we have as a DoD standard is internet protocol IP. The network protocol in a packet radio network which is the prototype example I'll be using throughout my talk is called CAP. At the link level we use something called carrier sense multiple access. This is a way of getting through on a particular link when neighboring nodes are sharing the channel. At the bottom level we have spread spectrum. We start out with an application called mail, the packet gets encapsulated in these various protocols, gets transmitted over to the next node using spread spectrum, the link level protocol makes sure that you can get across, the network level protocol tells you where to go next and you work your way through.

This model leaves out one thing, i.e., the control. Figure 4 The next thing we really ought to talk about is what I've dubbed the dual layered architecture. It is layering of the data-flow, and layering of the control architectures that support that data-flow. For example, at the physical level, spread spectrum, you have to control that physical media somehow. At the link level, carrier-sense multiple access, you have to have some control, understanding who your neighbors are so you know who to communicate with. At the network level you have to have packets that flow throughout the network controlling the network, establishing routes, flow-control, congestion-control, etc. The internet layer needs a similar requirement. To make that a little clearer, let me use some examples. Figure 5

If you recall, the physical level down here was spread spectrum, one example of physical control that has to occur is synchronization. The synchronization doesn't occur independently of the spread spectrum signalling it occurs on top of it. It makes use of the spread spectrum signaling to acquire synchronization. But the synchronization isn't part of the data-flow, it's part of the control of the link of the physical media. Similarly, neighbor tables are used to control the carrier-sense multiple access. A protocol we've got intra-network within packet radio, called SPP station to packet radio protocol makes use of the network protocol to do the control of the network itself. The data doesn't flow via SPP, this is all control packets. GGP gateway to gateway protocol makes use of IP to talk to the various gateways. That's all nice and theoretical. Now let me get to the subject of the talk.

The packet radio and what we tried to do in packet radio successfully (since the program is ending this year), is to combine the packet switching technology that we demonstrated initially in the ARPANET with wideband broadcast radio technology. Figure 6 Wideband meaning spread spectrum, broadcast meaning omni-directional. What a packet radio network is is a set of nodes, all identical. Figure 7 Each of these little circles is meant to indicate a packet radio with its top half the radio unit, the bottom half the digital controller, and what you do is you would attach devices to packet radios and you can attach terminal devices (not terminal=keyboard, but terminal=end). You can attach controlling elements called stations. You can attach host computers which are also terminal devices, or you can attach gateways to other networks. Sometimes you don't want to attach anything, you need a pure repeater because of the particular topology, for example, the particular geography you're working with prevents you from establishing connectivity between the groups of radios separated by a mountain. You cannot do anything with networking to overcome a mountain in the middle of your network. Packets flow through the network according to the routing protocols

etc., in each particular area the radios are all in connectivity, but this radio for example is not in connectivity with this one so it needs to use a repeater.

The management of that network with its mobile subscribers and broadcast capabilities is what the packet radio program was all about. In particular, Figure 8 the features we have in a packet radio are a packet switched store-and-forward network with spread spectrum radios Let me jump down here since the topic of the conference is spread spectrum, we use spread spectrum anti-multipath processing. Let me emphasize, what we have in the current packet radios and that's what I'll be getting to to make my point, is the current packet radio are spread spectrum but they're spread spectrum in an artificial way in that they have fixed codes. They have the same code for every bit and it's repeated ad nauseum. So it's spread spectrum in terms of waveform but it's not spread spectrum in terms of any sort of protection we normally think of in terms of AJ or LPI. You have to assume that the enemy is going to know your code since it's fixed, embedded in a SAWD. So this is a good way to look at what comes along as baggage with spread spectrum and all its associated control versus what you have when you have a basic network that's built up of everything except the actual spread spectrum capabilities. That's why I'm using packet radio as an example.

We are launching a new program called survivable radio networks Figure 9 and the idea behind survival radio networks is how do you take a network and make it robust in a large variety of threats and how do you have it handle very large networks like thousands of nodes. Obviously spread spectrum is going to play a key role here, otherwise I wouldn't be talking about it. What we are trying to do are two things. First of all we are going to be developing something called the low cost packet radio. Motivations of low cost radio was initially that in order to test a network of a thousand nodes, we had to buy a thousand radios and we couldn't afford a thousand high cost packet radios. So we needed low cost packet radios. But in typical DARPA fashion, rather than doing things simply we decided to make the problem a little more complicated and add in extra capabilities in the low cost packet radio. I'm glad that was done otherwise we wouldn't be able to have the program called survivable networks. Figure 10 The enhanced capabilities will permit experimentation with large robust networks. In particular, the thing we are going to be trying to do is develop the network management control strategies that allow you to manage a network with a high degree of robustness. How do you manage something like that ? How do you control the spread spectrum, etc. ? How do you tell when there's a threat that you need to respond to.

We've talked a lot about the low cost packet radio. Figure 11 This thing is about 400 cubic inches, costs about $10K, weighs about 12 lbs...nice radio. If you looked down these features here you will see most of the features are the same as the current packet radios for example, not low cost, but MSK modulation, L-band, 20-channels, bit-by-bit code changing is a significant difference. (I'll talk about that) but multipath accumulation, dual data, all the stuff is pretty much the same. We have a much more powerful microprocessor.

In particular, the additional features Figure 12 we are building in to the low cost packet radio are: bit-by-bit spread spectrum code changing under software control. That means that the software, the network level, gets to select what spreading sequence we use in every

packet. It can be based on time, it can be based on receiver, it can be based on some DES output. In fact what we really do is we take the output of the DES chip and use it as the spreading sequence, and what controls the sequence is the key i.e. the initialization vector, that goes into the DES chip. So we really have a lot of control there, pseudo-random codes. Another interesting property is that we've built in the capability to select the code sequence that gets used throughout the packet in the preamble of the packet. That means we can have a set (actually 8), different code sequences and in the preamble of the packet we get to choose one. How do you manage that ? How do you use it for CDMA ? How do you use it to establish virtual channel through the network ? These are the kinds of questions we have to address. The clock is useable for code sequence selection. What that means is it is accurate enough so that we can change codes at up to every 5 milliseconds and still maintain enough time synchronization so that we don't lose data at the boundaries of the guard times. But how do I maintain network time synchronization to do that ? Another control question. Forward error-correction, I'm really generating these codes at random and I've got a multiple-access environment. Occasionally you are going to have two users transmitting on the same code even though you are generating them randomly. There is a certain probability that that's going to happen. Forward error-correction is needed to take care of that, in addition to everything else that everybody has talked about. To do all this fancy network management we obviously need a powerful microprocessor.

Let me give you an example. Figure 13 This is what I call a time-slotted code operation. The code sequence is repeated in the preamble, just for synchronization, since this is only 28 bits, and then changes every bit throughout here. Among the parameters it is selected based on, is time, in particular, time-slot. As we move through this time line, if the packet started in the time-slot k we would use the code sequence that corresponds to that time slot. Since it's never the case that a particular radio transmits more than one packet in a given time-slot, what we effectively have is a completely non-repeating code from every radio, and if we allow the radio ID to be used in the selection of the code also, we have a completely non-repeating code through the network. Very powerful, but how do we manage it ? In particular, this is a set of technical issues. Figure 14 If you take a look down this set of technical issues, you will see that probably half of them are related to how do you manage a network that has the capability of changing the spreading keys in a networking environment ?

These are network management issues that we are trying to address. Figure 15 Although the statement of the problem is how do I design a large survivable network, half the issues wind up being, how do I manage a network that has the capability to do changing codes, spread spectrum. How do I manage a spread spectrum network ? Not how do I build a spread spectrum link, certainly we can do that. In fact, although we've chosen a particular spread spectrum technology here, all of these questions apply equally to any of the technologies that we've talked about. The bottom line, I think I'm trying to convince you of is that in terms of support for data, the actual data flow through the network, it is probably true that you can take a nice clean layered look at the architecture and say that you can build the network and have routing and all those kind of nice issues, and support data flow through the network very nicely, even if the links are spread spectrum. That's not the issue. The issue is how do you

control those spread spectrum links on a network-wide basis.

SPREAD SPECTRUM NETWORKS

BARRY LEINER DARPA/IPTO	LAYERING & SPREAD SPECTRUM
BUD GRAFF USA/CECOM	SPREAD SPECTRUM DATA NETWORKS
JOHN OLSEN HUGHES	NETWORKING IN A TACTICAL ENVIRONMENT
JOHN WOZENCRAFT NPG	TIME SLOT ALGORITHM FOR PACKET RADIO

FIGURE 1

Layered Architecture

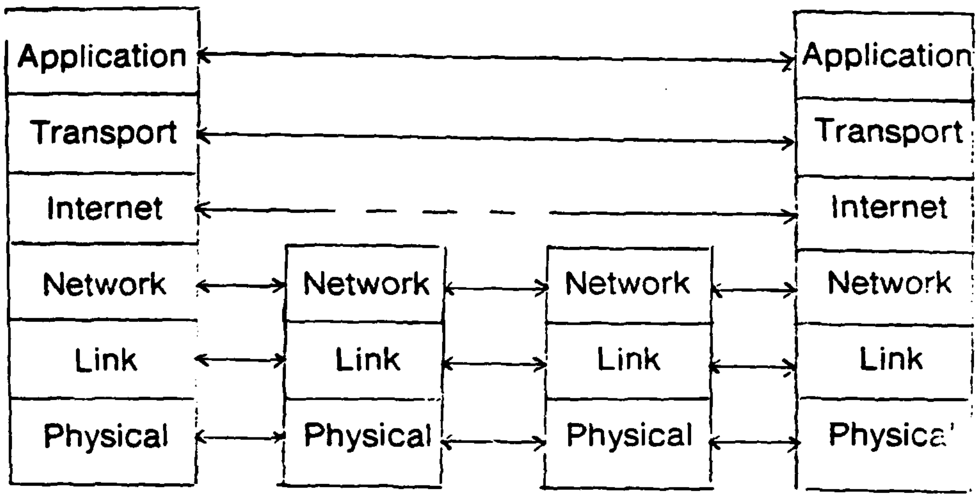


FIGURE 2

Layered Architecture (Example)

Application (Mail)			Application (Mail)
Transport (TCP)			Transport (TCP)
Internet (IP)			Internet (IP)
Network (CAP)	Network (CAP)	Network (CAP)	Network (CAP)
Link (CSMA)	Link (CSMA)	Link (CSMA)	Link (CSMA)
Physical (SS)	Physical (SS)	Physical (SS)	Physical (SS)

FIGURE 3

Dual Layered Architecture (Example)

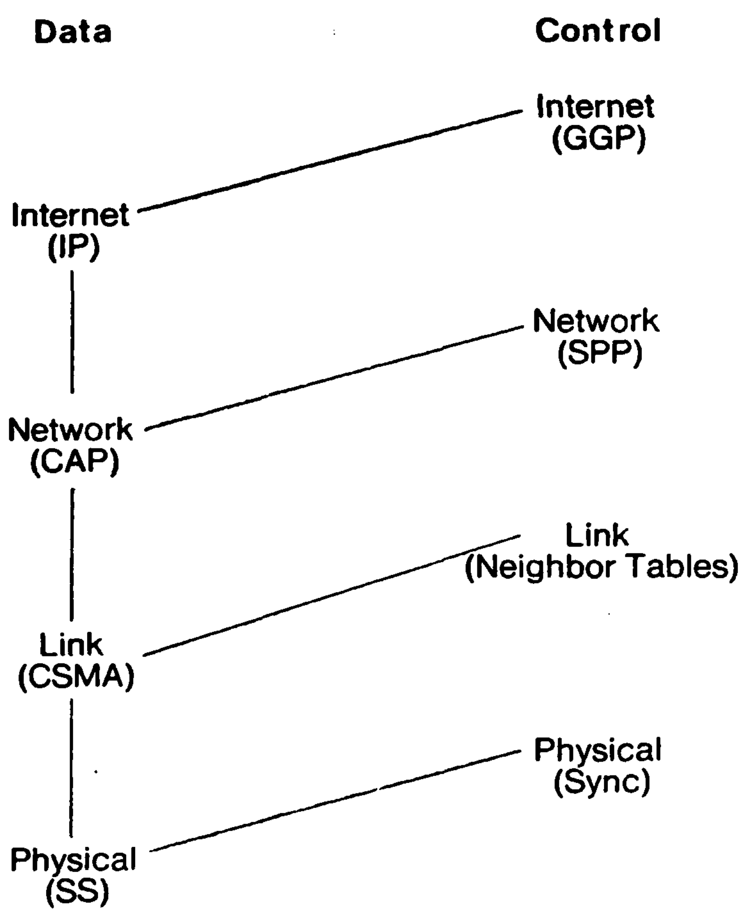


FIGURE 5

Dual Layered Architecture

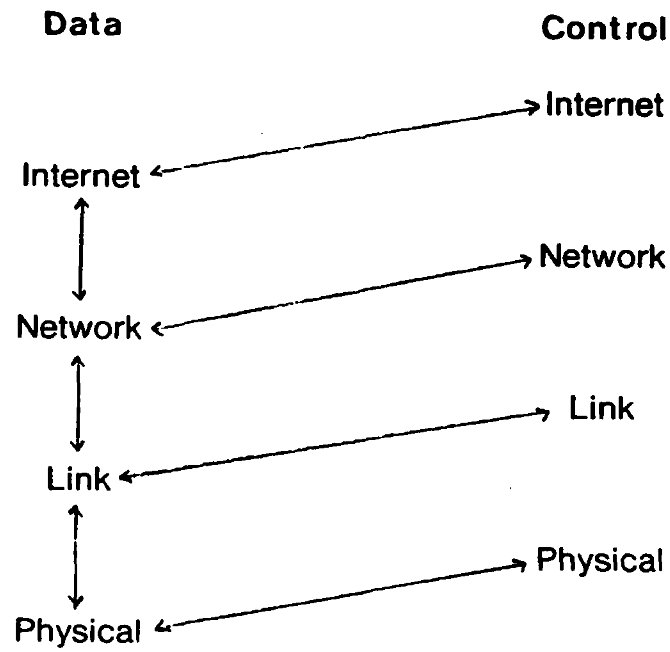


FIGURE 4

PACKET RADIO

A SYSTEM COMBINING
PACKET SWITCHED TECHNOLOGY
AND
WIDEBAND BROADCAST RADIO TECHNIQUES

FIGURE 6

PACKET RADIO NETWORK

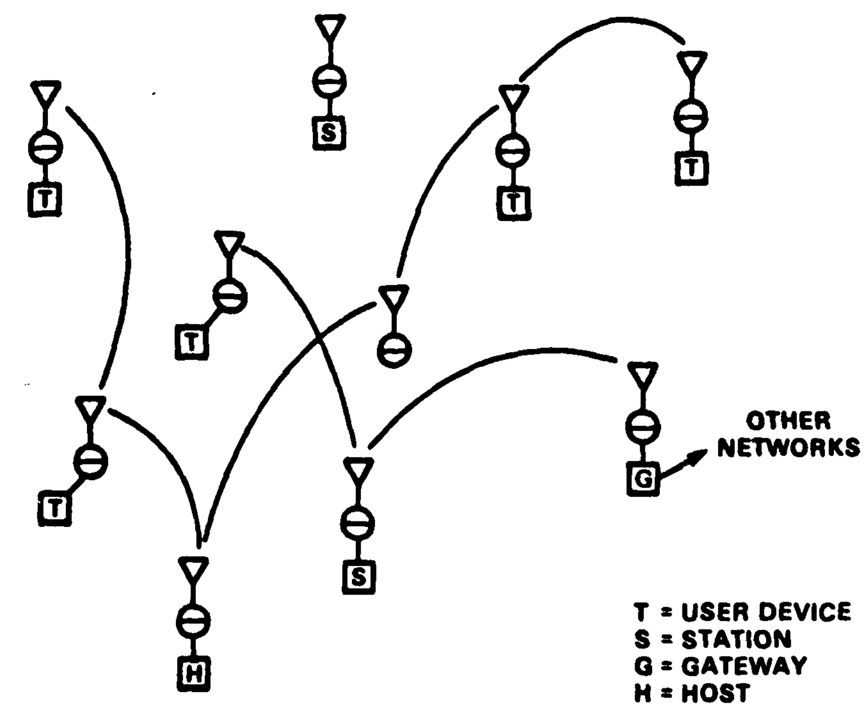


FIGURE 7

PACKET RADIO

System Features

- Packet switched store and forward network of spread spectrum radios
- Radio channel access control using Carrier Sense Multiple Access (CSMA) protocol
- Adaptive routing and network management and control
- Distributed/decentralized network control for survivability
- Internetwork capability
- Automatic transparent initialization and operation
- Spread spectrum anti-multipath processing

FIGURE 8

SUrvivable RAdio Networks

SURAN

FIGURE 9

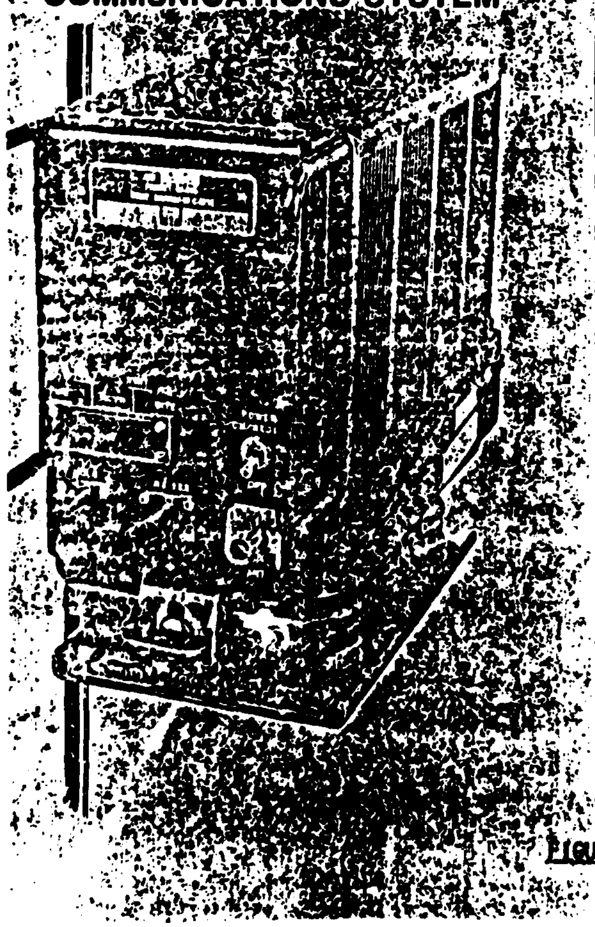
SURAN

PROGRAM OVERVIEW

- Develop and procure low-cost packet radio (LPR) with enhanced capabilities to permit experimentation with large survivable networks
- Develop and demonstrate network management and control algorithms to permit operation of a large network in the face of a variety of threats
- Investigate vulnerabilities of networks to sophisticated network-based countermeasures, and methodologies for eliminating these vulnerabilities

FIGURE 10

LOW-COST PACKET RADIO AN/PRC-118( )
COMMUNICATIONS SYSTEM



- LOW-COST
- 5000 HR MTBF
- MSK SPREAD SPECTRUM MODULATION/ DEMODULATION
- L-BAND: 5 WATTS (5 – 5 dB STEPS) –6 dBi OMNI ANTENNA
- 20 CHANNELS – SELECTABLE ON A PACKET-BY-PACKET-BASIS
- BIT-BY-BIT CODE CHANGE
- 12.8 M CHIP PN CODE
- MULTIPATH ACCUMULATOR
- DUAL DATA RATE: 100/400 kbps
- 28-BIT PREAMBLE (100 kbps FIXED RATE)
- 8086 MICROPROCESSOR
- MEMORY: 48k RAM, 16k ROM (WORDS)
- DUAL 8089 I/O PROCESSORS
- HDLC INTERFACE: CCIT X 0.25 LAP B, RS-449/422
- RS-232 INTERFACE
- FEC (FORWARD ERROR CORRECTION) – RATE 1, 7/8, 3/4, 1/2
- 10^{-12} UNDETECTED BER (CRC)
- MODULAR CONSTRUCTION FOR FUNCTIONAL EXPANSION
- POWER: 22 TO 32 VDC; 60 WATTS
- WEIGHT: 15 LB (6.8 kg)
- SIZE: 8.6" (21.9 cm) D,
  5.7" (14.6 cm) W,
  8.5" (21.6 cm) H

FIGURE 11

Low Cost Packet Radio

Additional Functionality

- Bit by bit spread spectrum code changing under software control (DES based)
- Ability to select code sequence in preamble
- Clock usable for code sequence selection
- Forward error correction
- More powerful microprocessor with additional memory

FIGURE 12

Code Changing Timeline

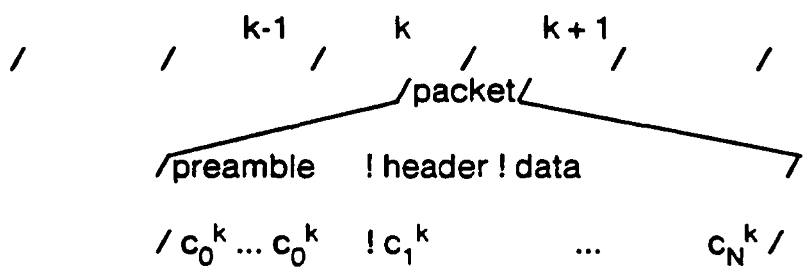


FIGURE 13

SURAN

TECHNICAL ISSUES

- Development of low-cost packet radio having bit-by-bit code changing, forward error correction and enhanced processing capability
- Management of large mobile network with distributed control
- Management of network with spread spectrum code division multiple access capability
- Protection of network management and control
- Dissemination and updating of spread spectrum code keys
- Operation in a threat (jamming, node capture, etc.) environment
- Operation in a radio silent or LPI mode
- C3 countermeasures based on networking technologies

FIGURE 14

SURAN Net Management Issues

- Management techniques for very large networks (1000 nodes) with large fraction of mobile nodes
- Survivability and graceful degradation
- Code slotting/time synchronization
- Key distribution and maintenance
- Code division multiple access operation
- Protection of network control
- Distributed support functions (e.g. name servers, network monitoring)
- Threat detection and responses
- Type of Service

FIGURE 15

CHARLES GRAFF

Viewgraph 1 I'm Charles J. Graff, from the US Army, CECOM, CENCOMS. The title of my talk is Spread Spectrum Data Networks. I'll be talking primarily about real-time data networks as opposed to spread spectrum technology. That has implications that I think you'll see as my talk progresses. First, a bit of motivation and background as to why the Army is interested in this topic.

Viewgraph 2 TRADOC has evolved a concept of tactical warfare called Air Land Battle 2000, which describes the way the US forces may fight in the year 2000 and beyond. There are some very interesting technical challenges to the communication community as well as giving us a concept of tactical operations.

Viewgraph 3 Basically the concept indicates a large number of users. They don't explicitly say multi-hop, but they do say spread out to survive. That implies to me, you have an arbitrary topology, and you can't site relays as you desire. The network connectivity, that is who can hear whom, must be assumed to be a strong function of time because user terminals are in motion, and you want to be able to communicate while you are on the move. Relays or terminals themselves may be out of operation or down for one reason or other, and you have to pay attention to jamming the environment as well. These three factors contribute to a connectivity change in a network, and I'll explain the implications of that as I go along. I mentioned before the need to disperse or to spread out is now one of the critical features in the tactical operations. The implication of that is that you want to have a degree of survivability in a network control. The typical kind of picture that they show is a military unit of some type, separated by some distance from other units of the same force. Thus, the analogy is that the battle will be fought in a manner of soccer, rather than football. This creates a bit of a communication problem for us to work on.

Viewgraph 4 I should mention another issue that we really shouldn't forget about in any tactical system is that of security, and its implications. The requirement for access control in a military system, key-distribution for both the communication security as well as Transmission Security (TRANSEC), and of course the usual military orientation of light weight, low power, low cost.

Viewgraph 5 Barry Leiner talked about bit-error rates, and probability of successful transmission, and power etc., but from the user's viewpoint, he cares about these kinds of issues, as shown in Viewgraph 5. What is his end-to-end delay in getting his messages through? What is his probability of success, i.e., what is the probability that the link is up, or that the network is up? And of course there are also the usual issues of survivability and reliability. This is where the user is coming from, not the link level, or the bit-error rate level, or whatever. These kinds of questions are what the user is interested in.

Viewgraph 6 I have a similar chart to Barry's regarding the networks. since it is a networking seminar I have a little different view of the world. Everybody has their own version of what a network model should look like. I have two here. My version of the DARPA model has 7 instead of 6 layers. The ISO (International Standards Organization) is an organization that tries to provide some standardization in the area of communication networks and they've developed a 7 layer or level model. Basically it is a model similar to the DARPA model. What we are concerned about here primarily is the communication subnet, which basically is from transport layer down which includes these bottom 4 layers. I would like to show how dynamic

spread spectrum affects both the network layer and the transport layer to an extent. That's going to be the focus of my talk here.

Viewgraph 7 Just a little more detail will be given about each one of these layers. The physical layer is normally what one would think in terms of signalling, in terms of pulse rise time, fall time, modulation, that kind of thing. The link level basically has as its goal the integration of these various bits or signal units into a link or frame; channel access of various types - carrier sense multiple access (CSMA), ALOHA, is performed there, and with spread spectrum modulation you can perform code division multiple access (CDMA). The network layer which is above that performs the integration of these different links, which are spread spectrum controlled, and maybe having dynamic power control on them as well, into a network of a multi-hop environment. The inter-network, as Barry also mentioned, is the integration of these different kinds of networks into a network of networks. For example, you could have a spread spectrum network which uses code division multiple access or frequency hopping integrated with a point-to-point long haul network which is a backbone network, and by a proper protocol layering technique you could make that kind of a system operate. That's the advantage of a layered architecture, you can see what kind of function has to be done where, and partition the problem into a sequence of small subproblems. Finally, the uppermost communication layer, the transport layer, provides source and destination communication between basically user end processes in a computer. A process is simply thought of as a computer program that's running somewhere in one machine or microprocessor. The type of communication service provided by the transport layer includes the familiar IPC (or InterProcess Communication) between remote processes, and the more traditional message passing or electronic mail. This is the kind of communication environment that we are working with.

Viewgraph 8 I want to emphasize again that the driver for all of this is what the user sees at the end-to-end level, and that's reflected down from the transport to the internet layer, from internet down to network, from network down to the link layer and the physical layer. So it provides a top-down driven approach to the problem.

Viewgraph 9 You are trying to satisfy the user's requirements or needs as he perceives them. If you take all these things and look at what that means from a communication network point of view, you find you have a network of some arbitrary topology as shown here. The goal is to provide communication between all pairs of nodes as required for a multi-hop environment that is perhaps changing connectivity due to jamming or other conditions. An additional complication is that the actual requirements for traffic loads aren't that clear. He (the user) only has a feeling about what his requirements are, and that has implications toward how you design the system. Certainly nobody has a good understanding of the traffic patterns in the dispersed environment, which is what is implied by Air Land Battle 2000. Due to the changing battlefield connectivity, we need a dynamic control technique to control communications.

Viewgraph 10 The point of all this is that we have a networking problem and the technical approach from the R & D community is to look at two types of adaptive network control, with network management level and also adaptive link level. Network management as I'm calling it here, means that we are considering the dynamic environment to be a critical driver and, as Barry pointed out, we are interested in developing control studies for

particularly these kinds of layers (Network Layer and Link Layer).

Viewgraph 11 I'd like to talk a little bit now about what that really means. I will give a little quick view of the network layer. I would like to talk about one type of function performed in the network layer, i.e., routing. That's probably the easiest to understand.

Viewgraph 12 Think of what routing implies. You have to first figure out what the connectivity is, and the connectivity is determined by the link characteristics which are spread spectrum modulated and controlled. Once you figure out what the connectivity is, you usually do some kind of path computation. Basically this is to find a set of links that you can use to go from an arbitrary source to desired destination. Once the path is computed, each packet may be forwarded toward the destination with the complete route, or *only next* hop address. So these three phases are generically thought of as the routing process. I'll show how the link characteristics affect basically these two kinds of function (i.e., the connectivity assessment and the path computation) in the routing layer.

Viewgraph 13 Typically, you can think of the one hop connectivity of users in a system as stored in a matrix form. An entry here could mean that this user can talk to this user here, or user i can directly communicate with user j. There are algorithms which are well known and well understood about how to find paths through the network based upon any criteria you want; from single source to all destinations is known as Dijkstra's algorithm of minimum cost; you can use an "all pairs" algorithm to find the minimum cost links between all pairs. You can minimize utilization in the highest loaded link as another optimization criteria. There are many different possible network parameters you could base a routing decision on. The point of "almost all" means that any routing algorithm will find the path or several paths between two arbitrary network users, provided that a path exists. The question is if there isn't a path, is there some way to make a path by slightly altering the connectivity?

Viewgraph 14 I mentioned before the connectivity matrix; the viewgraph shown here describes specifically what's in a typical connectivity matrix. Basically you could have a Boolean variable, meaning *link or no link, which is the simplest representation.* You could have a delay estimate of what the traffic delay would be using that particular link. There is another measure, traffic intensity or λ, a throughput estimate basically, in terms of keeping the link busy or heavily utilized. Or there's some multi-dimensional routing parameter which combines several other functions into some kind of routing metric. Thus, the spread spectrum link, with its link performance of delay, throughput, and error probability, will have a direct impact on the routing metric.

Viewgraph 15 I want to mention a little bit about distributed control of the spread spectrum link because it has an impact on the topology. Think of a network as something very simple like this in the upper diagram. Two nodes, i and j, are communicating, and each one of them has a full routing matrix. Typically the way the routing process works is that you have to send updates telling what the topology is between these two nodes which can communicate. Obviously, if node i is three hops away from node j, the information received is old by three hop times or three update periods. So there's a potential instability in terms of the routing algorithm because the information that this node (i) sees may be old because the connectivity may have changed, on the extreme left or right side of the network. Just suppose that you had an adaptive

spread spectrum link which you could control the power on, or the adaptive anti-jamming (A-J) on the link; you could maybe collapse or expand this kind of a network (upper figure) into this network (lower figure); you now have a lot of one-hop neighbors, which make the routing problem a little easier because now you are one-hop away. But it creates other problems in the sense that now you have to keep bigger tables, and some systems you can lose throughput due to packet collisions, because these links may interfere with each other. So it's not obvious how to find an optional solution to this kind of problem. The point of the matter is that, with an adaptive spread spectrum system, you do have a potential for coverting from this kind of picture (upper) to this kind of picture (lower).

Viewgraph 16 Here are some of the adjustable parameters you might be able to adjust on a link by link basis. You could adjust the power if you can figure out how to do that optimally in a network sense. You could adjust the antenna gain or directivity and you could do some tradeoffs in data rate/A-J performance. You could also do something with the coding gain. The point of the matter is you want to optimize the link performance on a network wide sense rather than on a one-hop or link basis. That's kind of where we are hoping to go.

Viewgraph 17 Given that you have this one-hop path from A to B, it's not obvious in a jamming environment, a stressed environment, that the link is symmetric, that is A to B or B to A have the same kind of characteristics. So there's a potential for some kind of control that has to be passed around the system. But, on the other hand, if you are only sending traffic from A to B and A originates the traffic, his routing decision is fairly simple. He just knows that B is one-hop away from him. It may be a long hop in terms of delay but it's basically one-hop. However, a node over here which is more than one-hop away has not very good knowledge about this link. So the question is how do you organize these link type parameters and variables into something that makes sense from a network level where the routing is done. And that's a question I don't have any answers for.

Viewgraph 18 The other possibility of course, is if you do have some control of this spread spectrum link in terms of power and A-J you may want to generate more one-hop neighbors, in the sense of using a directive antenna to pick up this node (c) as opposed to not hearing him at all, and that certainly has an advantage in terms of obtaining a richer connectivity. All these parameters are link level, spread spectrum control which you may be able to control somehow and they all impact on the routing and the connectivity matrix. That's basically the thrust of my talk here.

Viewgraph 19 If you do have this adaptive spread spectrum link control, the connectivity matrix is somewhat under network management in terms of adapting the topology slightly to suit your needs. You may or may not have two-way links. You may have the opportunity to use independent parallel paths for source to destination and/or different reverse paths. The connectivity is also a function of time, as I mentioned before, in terms of user-motion, of terminals and obstruction of terminals in the jamming environment, or whatever. Whenever you have adaptive routing algorithms, you have to be concerned with the stability, and what is the situation for steady state; or will there ever be a steady state in a very dynamic environment. These are some of the questions I see in terms of adaptive spread spectrum link control as it appears at the network level.

Viewgraph 20 Here is a very brief one-page summary. We need a better

understanding of adaptive distributed algorithms for basically these kinds of things as listed on this viewgraph. For example, the preamble control, key distribution for spreading codes, and an assignment of paths for a relay. You have to generate some kind of a link metric that makes sense to the network layer for a spread spectrum system, as contrasted with the simple delay metric often used today. You do have the ability now by having some control over the link level, to make the topology somewhat adaptive to your control, if you can figure out how to do that in some optimal sense. For example, an articulation point is a node in a network, which if you lose that node, you have the network now as two separate pieces and can't communicate. Such a network is not very survivable. We know of algorithms that will add additional links to enhance survivability provided you make the link physically occur there, either by increasing power or using directional antennas. Similarly with bifurcated routing, if you want to generate independent parallel paths for survivability, you may be able to add additional links, so that parallel paths result. And of course, the final issue is performance. After you've done all this adaptive control, does the increase in performance justify the complexity in hardware/software? These are only a few of the questions we hope to try to address.

Thank you for your attention.

VIEWGRAPH 1

SPREAD SPECTRUM
DATA NETWORKS

BY

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CENTER FOR COMMUNICATION SYSTEMS (CENCOMS)

PRESENTED AT ARO WORKSHOP ON SPREAD SPECTRUM, MAY 1983

VIEWGRAPH 2

ARMY SPREAD SPECTRUM NETWORK

PROBLEM

AIR LAND BATTLE (ALB) 2000

VIEWGRAPH 3

ALB 2000

- LARGE No. OF USER NODES
- MULTI-HOP ENVIRONMENT WITH ARBITRARY TOPOLOGY
- DYNAMIC NET CONNECTIVITY DUE TO
  - TERMINAL/USER MOTION
  - RELAY OUTAGE
  - JAMMING ENVIRONMENT
- NEED TO BE DISPERSED IN OPERATIONS
- SURVIVABLE NETWORK CONTROL

VIEWGRAPH 4

OTHER ARMY ISSUES

- SECURITY
- ACCESS CONTROL
- KEY DISTRIBUTION (SECURITY/TRANSEC)
- MIL-ORIENTATION - LIGHTWEIGHT/LOW PWR/LOW COST

VIEWGRAPH 5

FROM END-USER VIEW

- END-TO-END DELAY
- THROUGHPUT
- $P_{success}$
- SURVIVABILITY/RELIABILITY

VIEWGRAPH 6

NETWORK MODELS

ISO	DARPA
APPLICATION	APPLICATION
PRESENTATION	UTILITY
SESSION	TRANSPORT
TRANSPORT	INTERNET
NETWORK	NETWORK
LINK	LINK
PHYSICAL	PHYSICAL

VIEWGRAPH 7

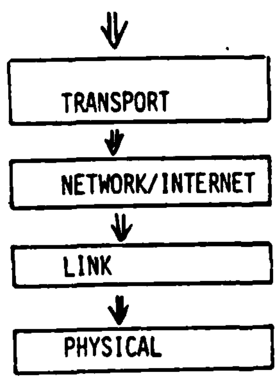
PHYSICAL	-	SIGNALLING
LINK	-	FRAMING/CHANNEL ACCESS (CDMA)
NETWORK	-	COLLECTION OF LINKS ⇒ NETWORK
INTERNET	-	COLLECTION OF NETWORKS
TRANSPORT	-	END USER COMMUNICATION

- PROCESS TO PROCESS COMMUNICATION (IPC)
- MSG ORIENTATION

VIEWGRAPH 8

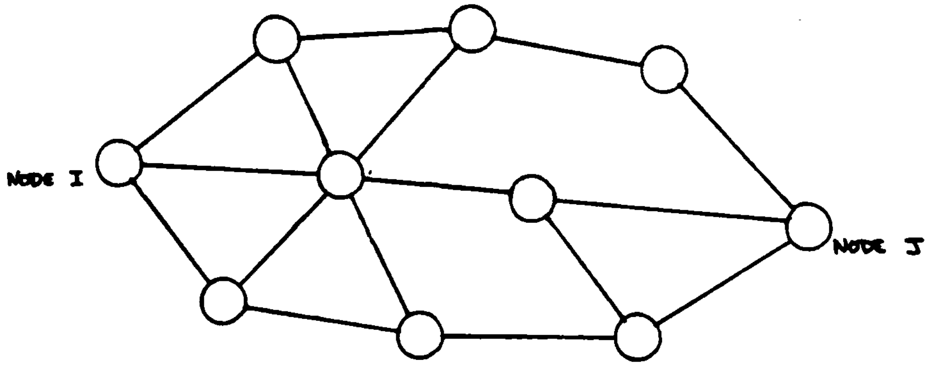
DRIVER IS

ETE USER COMM



VIEWGRAPH 9

NETWORK PROBLEM



GOAL: TO PROVIDE ETE USER COMMUNICATION IN DYNAMIC, HOSTILE, ENVIRONMENT; BUT USER TRAFFIC REQUIREMENTS R_{IJ} MAY NOT BE KNOWN OR WELL UNDERSTOOD FOR DISPERSED ENVIRONMENT.

VIEWGRAPH 10

TECHNICAL APPROACH TO NETWORKING

- ADAPTIVE NETWORK MANAGEMENT
- ADAPTIVE "LINK NET MANAGEMENT"

VIEWGRAPH 11

NETWORK LAYER

(E.G. ROUTING)

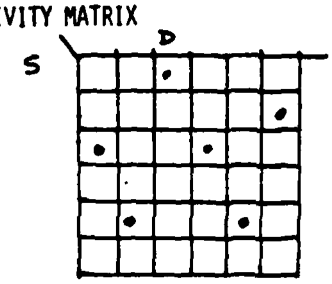
VIEWGRAPH 12

ROUTING ALGORITHM PHASES

- CONNECTIVITY ASSESSMENT
- PATH COMPUTATION
- FORWARDING

VIEWGRAPH 13

CONNECTIVITY MATRIX



TYPICAL PATH FINDERS

- SINGLE SOURCE - DIJKSTRA ALGORITHM (MIN COST)
- ALL PAIRS -
- MINIMIZE UTILIZATION OF HIGHEST ρ LINK

"ALMOST ALL" ALGORITHMS WILL FIND A PATH FROM A TO B IF ONE EXISTS

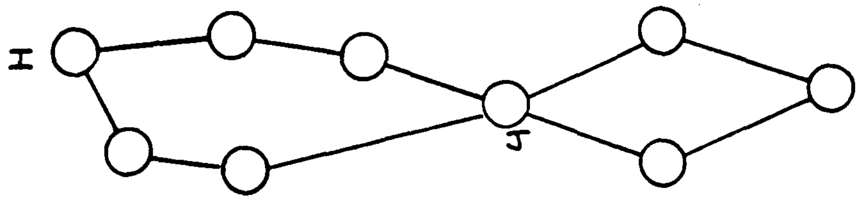
VIEWGRAPH 14

TYPICAL CONNECTIVITY MATRIX INFORMATION

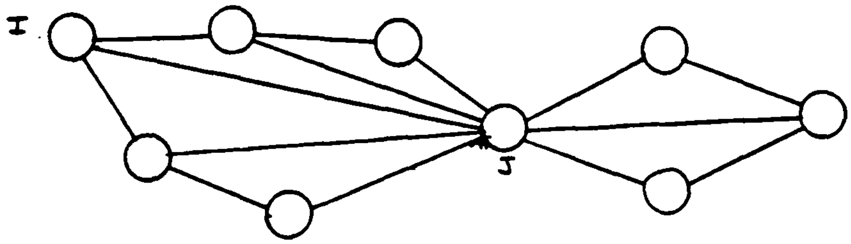
1. BOOLEAN (LINK-NO LINK)
2. DELAY ESTIMATE (SEC)
3. THROUGHPUT ESTIMATE (BPS)
4. MULTIPLE DIMENSION (PWR, P_E,....)

VIEWGRAPH 15

INFORMATION "AGE"   HOP DISTANCE FOR DISTRIBUTED ALGORITHMS



IJ - THREE HOPS   (OLD DATA)



IJ - ONE HOP   (FRESH DATA)

INSTABILITIES, INCORRECT/INCONSISTENT DATA IN "DISTRIBUTED ENVIRONMENT"

VIEWGRAPH 16

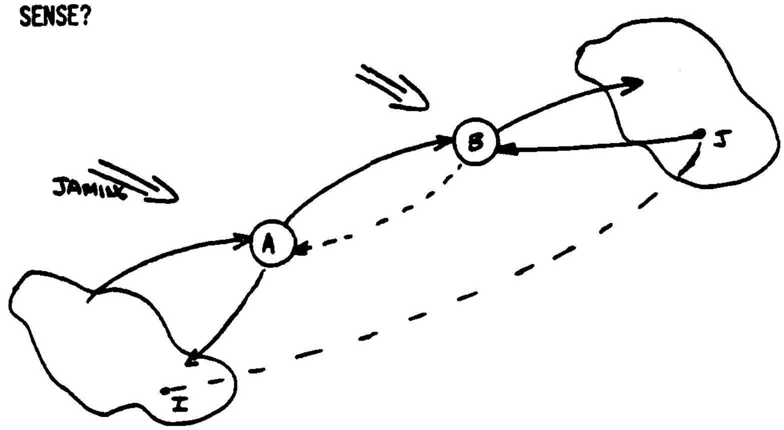
POSSIBLE PARAMETERS ADJUSTABLE ON LINK BASIS

- POWER
- ANTENNA GAIN/DIRECTIVITY
- DATA RATE - A/J
- CODING
- ETC.
-

ADJUST EACH ONE HOP LINK FOR "OPTIMUM" PERFORMANCE

VIEWGRAPH 17

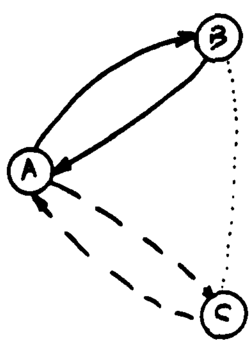
GIVEN THAT A LINK CAN BE MADE ADAPTIVE, HOW DO YOU CONTROL IT IN A NETWORK SENSE?



1. MAYBE $L(AB) \neq L(BA)$
2. $\lambda_{AB} \Rightarrow AB$ - ONE HOP FOR AB TRAFFIC
3. λ_{IJ} - AB HOP "OLD" INFORMATION

VIEWGRAPH 18

CREATION OF ADDITIONAL ONE HOP NEIGHBORS UNDER DYNAMIC LINK CONTROL



— BC INTERFERENCE MAY OR MAY NOT RESULT DEPENDENT ON SYSTEM CHARACTERISTICS

VIEWGRAPH 19

THEREFORE:

- CONNECTIVITY MATRIX IS SOMEWHAT UNDER NET MANAGEMENT CONTROL WITHIN LIMITS
- CONNECTIVITY MATRIX MAY NOT BE SYMMETRIC
  - ONE WAY LINKS
  - INDEPENDENT PARALLEL PATHS
  - DIFFERENT FORWARD/REVERSE PATHS
- CONNECTIVITY MATRIX IS FUNCTION OF TIME
  - MOTION OF TERMINALS
  - DESTRUCTION OF TERMINALS
  - JAMMING
  - USER TRAFFIC LOAD
- STABILITY OF ALGORITHM
- WILL THERE EVER BE A "STEADY STATE"?

VIEWGRAPH 20

TOPICS FOR RESEARCH

- ADAPTIVE, DISTRIBUTED ALGORITHMS
  - CHANNEL CONTROL (PWR, RATE, CODE, ETC.)
  - ALLOCATION/DEALLOCATION OF RESOURCES (PATHS/KEYS,....)
- GENERALIZED LINK METRIC USEFUL TO NETWORK LAYER
- "ADAPTIVE TOPOLOGY" VIA LINK CONTROL BIARTICULATION POINTS/BIFURCATED ROUTING
- PERFORMANCE ISSUES

BARRY LEINER: Our next speaker is John Olsen of Hughes Aircraft Company, Ground System Group. He is the Department manager in charge of the Advanced Data Links and Network Systems department. He received his Ph.D. from USC in 1978 under Bob Scholtz and he has spent 10 years at Hughes in anti-jam communications and data links with concentration in the last 4 years in anti-jam spread spectrum networks. And he'll be talking to us on networking in a tactical environment.

JOHN OLSEN: As Barry mentioned, what I'd like to talk about is networking in a modern tactical environment By a tactical environment I mean a ground radio communication network, a network which may contain thousands of radio terminals. These terminals are characterized as being in very irregular and rough terrain, having very low antennas in the terrain, and therefore can be expected to have very limited R.F. connectivity. Not only is the connectivity very limited by the rough terrain, but also by the fact the users are in constant motion in the irregular terrain, which causes the connectivity to change very radically and dynamically. On top of this we have also the effect of jammers, which will strive to limit, and change as a function of time, the connectivity as much as possible. In summary, there are two aspects of what happens to the connectivity which the net management has to deal with. First is that it's limited from the terrain masking in addition to the jamming itself. Second that it's changing very quickly.

Other factors that are very influential to the net management design in this tactical environment are: (1) you have users that are constantly turning on and off, and when a user turns on it must be accommodated into the network very quickly, and (2) you have very rapidly changing needs of the communication network's users in terms of their communication requirements. Any user may suddenly have a need to communicate with any other user in the network and that need has to be very quickly accommodated.

<u>Slide 1</u> Barry had mentioned that really the theme he wanted us to get to was what happens if you put spread spectrum and networking together, and why is the sum more than just a simple sum of the parts? My point of view is that, if you put spread spectrum and networks together, you get exactly that, a spread spectrum network. What I mean by that is with a network, if you just add a data link which is spread spectrum, what you will get is the same network with some improved communications capability, one that is in particular more immune to multipath. It has more reliable communications and it is less subjective to <u>non-intelligent</u> interference. That's the key point of it. What I want to say is that spread spectrum and net management will get you a spread spectrum network, but does not necessarily give you an antijam network.

The reason for that is, we are not dealing with just simple, white Gaussian noise at a high power level from the jammer. We have to deal with a very intelligent jammer that may try to spoof the network. Not only in terms of spread spectrum in locking up a single receiver, I'm talking about spoofing the entire network. When we develop a spread spectrum, or really I should say an anti-jam, communication network the premise that we have to operate on is that the enemy knows all the aspects of the system. We must assume he knows every detailed piece of information on the design of the system, including your spread spectrum waveform design, your PN code generator, the details of the error control coding, etc. We must assume the only

thing that's denied the enemy is the keys, the cryptographic keys for the system. In fact you'd have to even plan that, and accommodate into your network design and your network management design, very quickly the enemy will probably have a working unit of your equipment, meaning it will have a complete unit with a cryptographic devices at least intact, maybe not the keys.

What I want to say really is, to get to an anti-jam network there are really 3 components, and not 2. The first one is of course spread spectrum, that's necessary to have some links that will work in the environment. If you don't have spread spectrum communications you may not have any working links Secondly, is the automatic network management that provides all the relaying and resource allocation required in the network. The third item that is essential to put those two things together and obtain anti-jam is a *security architecture. However, this is* something that I really can't get into in very much detail here in this unclassified session.

Slide 2 I'd just like to say some words that will start you thinking of how important it is to have a security architecture supporting both the spread spectrum and networking to make an anti-jam communication network.

So what is a tactical net management concerned with? I think the concentration of research and certainly the more interesting aspects of net management (that have received by far the most research interest) are the network access protocols and the routing and relaying algorithms. But this is really only a small portion of the networking problem. One of the most important elements of net management is really the control and the monitoring aspect. How do you know how well your system is performing and how do you control it if it isn't working up to your expectations? A key aspect of *monitoring and control* is the management of the cryptographic net. This is the function that supplies you with a truly anti-jam network, including spread spectrum that works to what you anticipate it to be and also a network that works the way you expect it to in spite of the intelligent jammers. And to accomplish that, what gets into the network design is that you have a partition variable mode of operation. There are two aspects of this. One is what's called the traffic variable which provides transmission security (the name in the trade is called TRANSEC). What this allows the networker to do is have all the units available for relaying network information. That simply allows them *all* to be communicating with the same spread spectrum waveform. Although you generally want to be able to have all the users available as a relay of opportunity in the network, you do not necessarily want them to have access to your data. The message variable is what will supply the network with the actual protection of the information from source to destination because very few of the users have the need to know the information. That gets us into restriction of the access to the information by dividing up the network's users into what's called communities of interest, and there are certain requirements on size of these communities. By the way, one thing about restricting information that becomes key when we start looking at distributed network algorithms is to keep the control information off a global net that every user has access to. That's subject to be spoofable, certainly more so.

The control and monitoring aspect includes apparently simple but practical things that we have to be concerned with. For example, the counting of the security devices as these are classified pieces of equipment. They have to be accounted for at all times. Accounting for a security

device on a battlefield environment is a significant problem. In addition to the accounting for the devices is the accounting and also control of the access to the keys. That brings me to the question of how do you get the keys out there? You probably have to change them frequently because you may have one of your units captured. The solution is a network feature we call "over the air" rekeying i.e., you don't want to have to send someone out there to load the variables into units It's just impractical, just think what could happen if the key loading device were to be captured. You also must include as part of this monitoring function, how you detect if a compromise has been made. You have to have some means built in to your system to detect compromises and also to make corrective actions if there is a compromise suspected, like capture of a unit. In a suspected compromise situation there's not much you can do other than to rekey the net, with the exception of the terminal suspected of compromise. These are the types of things that are essential to really making the network practical so that you can actually put it out in the field. You have to get in and address these problems and have a sound, spread spectrum design, network management design and the security architecture that complements the other two components. I don't think I can really say much more than I've already said on this subject in this unclassified session.

My goal of introducing the security and monitoring aspects was only to give you a flavor of some really important and practical problems that are often overlooked. I think the spread spectrum technique aspect per se is pretty well covered. Also routing algorithms are in pretty good shape but it's really putting those items together and having a system that hangs together and really is secure in a hostile environment that is the key to developing anti-jam networks.

Slide 3 The next subject I would like to talk about is really changing the subject. That subject is, how do you characterize performance of a network of a spread spectrum network? If you have a simple data link, i.e., just a point-to-point spread spectrum transmitter and receiver, you can characterize that pretty well. You define a threat, you may have a pulse jammer at a certain rate, and you can go through some calculations knowing this spread spectrum waveform has certain parameters. The result of the analysis would be of the form that you'd expect that 50% message throughput at a certain signal to jammer power ratio, and then you can go to the lab and readily measure it. But when you get to a network, first off you have a significant problem even defining what is meant by performance. There are all different ways of characterizing performance. And secondly, how are you going to test network performance. Are you going to go out with your thousand terminals, go out in the field and tell everybody to do a certain thing and be able to control the exercise? We've found in the field testing we've done that there's no way of getting repeatable experiments. We have run small tests with 10-15 people. We tell them to go out, that they will see an X on the road, and to go there. They get out to the location and they look around ... "Hmmm, it's a hot day and there is a shade tree over there. I'm not going over to spot X, I'm going under the tree, it won't make a difference." And the whole test, the controllability of it is lost. By the way, simulations can frequently be used to characterize performance of large networks in a more controllable fashion.

The first problem I've mentioned is how do you define quantitatively, system performance measures. First off there are different performance measures that you

might come up with that make some sense. For example, the percentage of the communication requirements that are satisfied at a given time by meeting the throughput requirements and the response time or delay requirement. Something else very important is that if you have a major transient in terms of connectivity or user requirements onto the system, how quickly can the network adapt to accommodate the changes? That's a separate measure that you might come up with.

But what I want to get to really is what is the advantage of the adaptive networking? What does it buy you to have a spread spectrum network? How can you define what the advantage is if you had just a spread spectrum data link and now you add networking onto it? What is the networking worth in terms of should you go out and argue for 10 times more bandwidth to get more spread spectrum improvement, or is it automatic relaying buying more than increased *bandwidth or* increased transmitter power would do for you. If you had spread spectrum without networking and then added the networking on, but had the same threat, the same user requirements then how much more jammer ERP could the network handle for the same delivered performance? That would be a quantitative measure of the increased performance due to networking. For example, the network may be capable of operation in the face of 10 dB more jammer power and still meet the user requirements.

There was a paper in the 1980 Transactions on Communications, an individual from MITRE, a Mr. Cook came up with a way of characterizing the relative advantages of spread spectrum techniques and adaptive relaying. The title of the paper was "Optimum Employment of Communication Relays in an Interference Environment". If you specially wanted to communicate from the source to the destination as shown on the diagram, then what would be the minimum relaying required to accomplish communications with geometries shown for the user/interference environment. By the circles I've tried to indicate what is the communication region that is feasible for the receiver/transmitter with the given spread spectrum capabilities of the waveform. (The communications regions are not at all circles, but this point I will get to in the next chart.) Mr. Cook's paper gives you some very interesting insight into what is the advantage of spread spectrum with automatic relaying. But it's limited in that, to get some closed form type of results, he had to assume a certain geometry to the jammer, with the jammer and all the relays in a line, and he also assumed free space propagation loss, an r^2 type of loss which is obviously a somewhat dangerous assumption for characterizing a ground tactical communication network. However, some interesting results were *obtained*.

<u>Slide 4</u> The right type of results were obtained as you can see, this is a curve right out of Cook's paper. The factors on the abscissed, called the system survivability, here indicates what the spread spectrum is worth. It takes into account the ERP and antenna gains, bandwidths and processing gain, all those type of things. Then parametrically, is the different curves for how many relays you are allowed to have in the system, and then on the ordinate is what is the resultant system performance. You can see by looking across the curves that the networking is worth quite a bit. For example, if your system had no relaying capability but had a system survivability factor of 10, whereas if you had a one relay capability, you could reduce by 10 dB the processing gain and still obtain the same performance you had without the relaying. That type of improvement is really interesting.

I understand from Barry that he has also obtained similar results as to what's in this paper. The results are not surprising but it was refreshing to me that someone had tried to really quantify the performance advantage of relaying. Obviously it's limited to a certain application but it's a good start. I believe if you try to apply these results to a ground environment, I think that it will give a very limited and weak lower bound. The performance of a network is much better than that predicted by this bound, I'm sure.

Slide 5 Look for example at a network deployed on the ground with an air-borne jammer off in the distance trying to disrupt your network. Your units are very randomly dispersed on the ground with low antennas. With Cook's model, you simply try to come up with some sort of an average communication radius and not take into account the effects of the terrain, you get some sort of circular communications region. Again this region indicates that only the unit within the region you can communicate with. These average radius regions are typically very small and therefore it takes an extensive amount of relaying to reach the other side of the deployment. But in fact we find out that, and this is verified by field tests, you may have a communication region that looks something more like the highly irregularly shaped region on the diagram. You might justify this figure to yourself by noting that he's protected from the jammer behind a hill and that he could communicate with any user that's clearly shadowed by the hill. The key point is that even though the average radius might agree with that predicted by some of the models, there's many links that might work that are much greater in length than the average radius. Another factor of the real world is that you have so many available links to choose from and they are by no means constrained to fall along a straight line. There are so many units out there that there's often several units that provide links of significant length and in the right direction. You may have only a few directions where you can get out very far but there are so many units out there that there's bound to be one in the "good" directions. It turns out that there are frequently very many links that are available to use.

So this funny little pattern on the diagram is trying to show what an actual communication region might be. In fact, we have seen in the field tests, that in fact the network linking does look like that. These systems typically have a display where we can actually see what the network linking looks like in real time. You might be at first surprised by the way the linking goes. It doesn't necessarily go in what might seem like the shortest path. The networking utilizes links that are actually working and not what you might apriori think. So it may go all over the place to get from a network source to a destination. I think in all of the field tests that they've done against deployed jammers, the terrain has shown a very significant effect in shadowing these ground users and really giving an increase in system performance.

Slide 5 This brings me to some conclusion of what might be some topics for unclassified research, however, I'm afraid maybe some of the areas I have on my list could only be questioned in sufficient depth in classified research. These research topics, however, are all very relevant to anti-jam communication networks.

I just talked about how does one characterize network performance. That seems to me something that could be held in an unclassified environment. As long as you keep away from the exact characteristics of specific system or specific threat you are O.K., you can usually be safe if you talk in generalities.

You actually can use quantitative numbers as long as it is not tied into the capabilities of any system or any defined threat.

But a key element of quantification of the anti-jam advantage of networks, in fact the second area that Bud Graff started to get into in his talk, is the dynamic network performance. Say you have a large community containing thousands of terminals out there and it's come to a steady state. Then you have a major connectivity transient, for example the jammer just turns on, and what does it do to your network. I don't think that in terms of these large networks anyone's has really quantified dynamic performance.

Another important point is the capacity of the network, especially spread spectrum networks. With spread spectrum networks you have several different available dimensions, like the code dimensions, the frequency dimension in addition there is another dimension that people may not think of which is the space dimension. When we have a large geographical network, you can have different units using the same frequency/code space simultaneously, and they won't interfere with each other. There is a big capacity reservoir here. So quantification of what is really the capacity, in terms of where does the system break down, can be quite difficult. You can put so much source traffic into the network, and the question is when does the delay start going up significantly? This should prove to be an excellent unclassified research topic.

I'll mention again that the problem that really binds us in trying to come up with quantitative performance characterizations of the ground networks is the propagation modeling. The models just don't support the type of analysis we need to perform. There are two aspects of this. One is the absolute level of propagation loss at a given point in time. The second thing is how quickly is it changing. That's why I'm concerned with what I call spatial correlation. If the unit moves, how quickly does its propagation loss change, it is a very key aspect for modeling how quickly your network's links are switching. In addition there is foliage loss, a whole other realm. The ground based units are right out there in the trees, and that's a big propagation problem at any of the frequencies that we've been talking about here.

The next item here, development of distributed net control algorithms, at least in my view, there's been a lot of good work done here. However, I believe one significant problem with distributed control is trying to get the security aspect into it. One of the things about distributed algorithms is that they rely on everyone having access to the network control information, and that's dangerous in terms of being subjectable to *spoofing*.

Again I would like to suggest that emphasis needs to be placed on the control and the monitoring aspects of net management.

The last item I have, something that has gotten very important to us, but may not at first sound like it's particularly communication-related, is network multi-level security. If you have a terminal that has two different data sources coming into it, for instance one may be an unclassified source while the other may be a secret source, how do you handle the segregating and separating of the data so that the data classification levels are not violated? So you have to come up with designs that support a multilevel security network architecture. What the future holds as a solution to multilevel security problems is what is called trusted software. That might appear to be the strangest thing to have on an unclassified research topics list, but I think that most

of the work that's been done on trusted software turns out to be unclassified. Previously this work was under the auspices of the National Security Agency, and I believe it's very recently been effectively moved out of NSA into another organization of the Department of Defense called the Computer security center. I understand the intent of this separate organization is to emphasize that they really want to reach and interact with the technical community on the subject of trusted software.

NETWORKING IN THE MODERN TACTICAL ENVIRONMENT

HUGHES

- SPREAD SPECTRUM + NET MANAGEMENT $\nRightarrow$ ANTI JAM NETWORK
  - INTELLIGENT ADVISARY
  - ENEMY IS ONLY DENIED THE "KEYS", WILL ATTEMPT TO SPOOF NETWORK
- ANTI-JAM COMMUNICATIONS NETWORK HAS 3 CRITICAL ELEMENTS
  - SPREAD SPECTRUM
  - NETWORK MANAGEMENT
  - SECURITY ARCHITECTURE

SLIDE 1

TACTICAL NETWORK MANAGEMENT

HUGHES

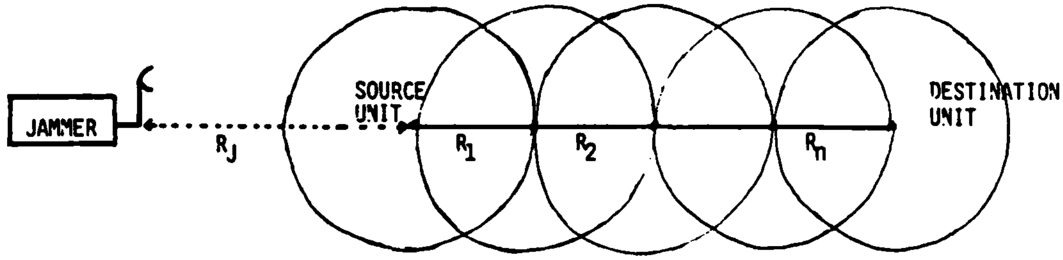
- NETWORK ACCESS PROTOCOLS
- ROUTING/RELAYING ALGORITHMS
- NETWORK CONTROL AND MONITORING
- MANAGEMENT OF CRYPTO NET
  - PARTITION VARIABLE OPERATION
    - TRAFFIC VARIABLE (TRANSEC)
    - MESSAGE VARIABLE (MSEC)
  - RESTRICTION OF ACCESS TO INFORMATION (COMMUNITIES OF INTEREST)
    - KEEP NETWORK CONTROL/MONITORING DATA OFF "GLOBAL" NET
  - ACCOUNTING FOR SECURITY DEVICES
  - ACCOUNTING/CONTROL OF ACCESS TO KEYS (i.e., OVER THE AIR REKEYING, OTAR)
  - DETECTION OF COMPROMISE (e.g., CAPTURE OF TERMINAL)
  - CORRECTION OF COMPROMISING SITUATION

SLIDE 2

QUANTIFICATION OF ANTI-JAM ADVANTAGE OF NETWORKING

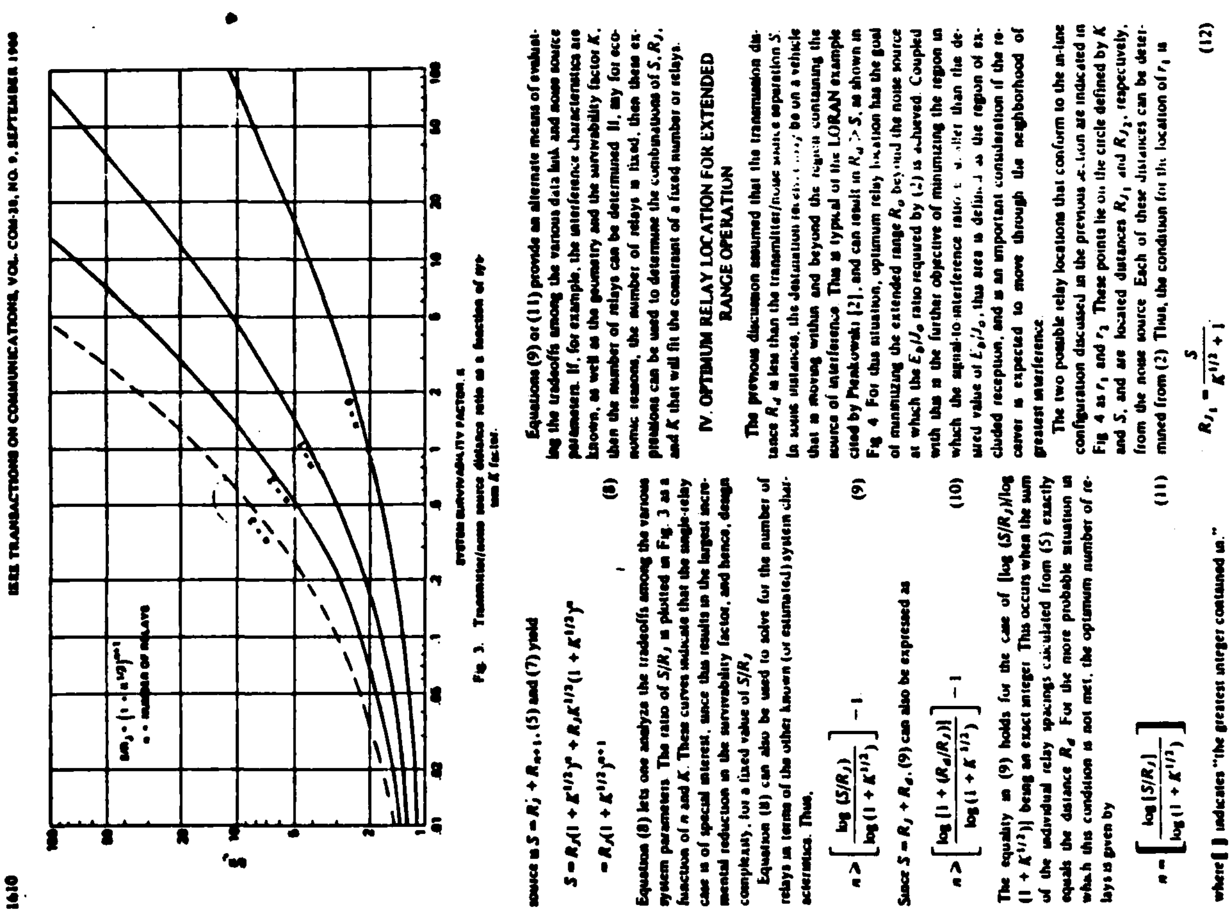
HUGHES

- FIRST PROBLEM IS TO DEFINE QUANTITATIVE SYSTEM PERFORMANCE MEASURE
  - e.g., % OF COMMUNICATIONS REQUIREMENTS SATISFIED (THRU-PUT, RESPONSE TIME/DELAY)
  - e.g., RESPONSE TIME FOR NETWORK TO RECOVER FROM MAJOR JAMMING TRANSIENT
- THEN COULD DEFINE AJ ADVANTAGE AS THE INCREASE IN JAMMER ERP THAT COULD BE TOLERATED FOR THE SAME PERFORMANCE AS ACHIEVED WITHOUT NETWORKING.
- NETWORK AJ QUANTIFICATION APPROACH OF R.E. COOK*(MITRE - BOSTON)



*"OPTIMUM DEPLOYMENT OF COMMUNICATIONS RELAYS IN AN INTERFERENCE ENVIRONMENT", IEEE TRANSACTIONS ON COMMUNICATIONS, VOLUME COM-28, NO. 9, SEPTEMBER 1980.

SLIDE 3



1610 IEEE TRANSACTIONS ON COMMUNICATIONS, VOL. COM-28, NO. 9, SEPTEMBER 1980

Fig. 3. Transmitter/noise source distance ratio as a function of system K factor.

source is $S = R_J + R_{n+1}$, (5) and (7) yield

$$S = R_J(1 + K^{1/2})^n + R_J K^{1/2}(1 + K^{1/2})^n = R_J(1 + K^{1/2})^{n+1} \quad (8)$$

Equation (8) lets one analyze the tradeoffs among the various system parameters. The ratio of S/R_J is plotted in Fig. 3 as a function of n and K. These curves indicate that the single-relay case is of special interest, since this results in the largest incremental reduction in the survivability factor, and hence, design complexity, for a fixed value of S/R_J.

Equation (8) can also be used to solve for the number of relays in terms of the other known (or estimated) system characteristics. Thus,

$$n > \left[\frac{\log (S/R_J)}{\log (1 + K^{1/2})}\right] - 1. \quad (9)$$

Since $S = R_J + R_d$, (9) can also be expressed as

$$n > \left[\frac{\log [1 + (R_d/R_J)]}{\log (1 + K^{1/2})}\right] - 1 \quad (10)$$

The equality in (9) holds for the case of $[\log (S/R_J)/\log (1 + K^{1/2})]$ being an exact integer. This occurs when the sum of the individual relay spacings calculated from (5) exactly equals the distance R_d. For the more probable situation in which this condition is not met, the optimum number of relays is given by

$$n = \left[\frac{\log (S/R_J)}{\log (1 + K^{1/2})}\right] \quad (11)$$

where [ ] indicates "the greatest integer contained in."

Equations (9) or (11) provide an alternate means of evaluating the tradeoffs among the various data link and noise source parameters. If, for example, the interference characteristics are known, as well as the geometry and the survivability factor K, then the number of relays can be determined. If, for economic reasons, the number of relays is fixed, then these expressions can be used to determine the combinations of S, R_J, and K that will fit the constraint of a fixed number of relays.

IV. OPTIMUM RELAY LOCATION FOR EXTENDED RANGE OPERATION

The previous discussion assumed that the transmission distance R_d is less than the transmitter/noise source separation S. In some instances, the destination receiver may be on a vehicle that is moving within and beyond the region containing the source of interference. This is typical of the LORAN example cited by Pienkowski [2], and can result in $R_d > S$, as shown in Fig. 4. For this situation, optimum relay location has the goal of minimizing the extended range R_o beyond the noise source at which the E_b/J_o ratio required by (2) is achieved. Coupled with this is the further objective of minimizing the region in which the signal-to-interference ratio is smaller than the desired value of E_b/J_o; this area is defined as the region of excluded reception, and is an important consideration if the receiver is expected to move through the neighborhood of greatest interference.

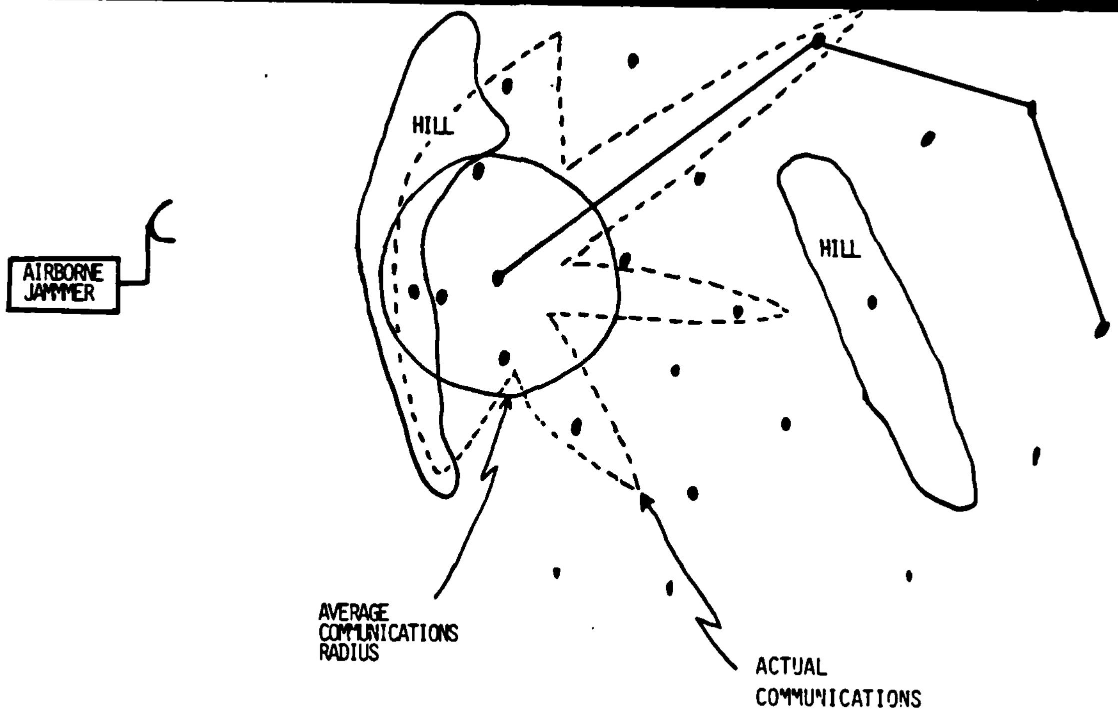
The two possible relay locations that conform to the in-line configuration discussed in the previous section are indicated in Fig. 4 as r_1 and r_2. These points lie on the circle defined by K and S, and are located distances R_{J_1} and R_{J_2}, respectively, from the noise source. Each of these distances can be determined from (2). Thus, the condition for the location of r_1 is

$$R_{J_1} = \frac{S}{K^{1/2} + 1} \quad (12)$$

SLIDE 4

ANTI-JAM ADVANTAGE OF NETWORKING IN A GROUND ENVIRONMENT IS MORE DIFFICULT TO QUANTIFY

HUGHES



SLIDE 5

TOPICS FOR UNCLASSIFIED RESEARCH RELEVANT TO AJ COMMUNICATIONS NETWORKS

HUGHES

- NETWORK PERFORMANCE
  - QUANTIFICATION OF ANTI-JAM ADVANTAGE OF NETWORKING
  - DYNAMIC NETWORK PERFORMANCE
  - CAPACITY OF NETWORK (SPACE REUSE FACTOR)
- PROPAGATION MODELING
  - SPATIAL CORRELATION
  - FOLIAGE LOSS
- DEVELOPMENT OF DISTRIBUTED NET CONTROL ALGORITHMS WHICH ARE NON-SPOOFABLE
- CONTROL AND MONITORING ASPECTS OF NET MANAGEMENT
- NETWORK SECURITY
  - MULTI-LEVEL SECURITY NETWORK ARCHITECTURES
  - "TRUSTED" SOFTWARE

SLIDE 6

BARRY LEINER: My understanding is that the C-Groups, the computer security group that John just alluded to, is part of NSA but they have a charter to make sure that they communicate better with the rest of the world than NSA traditionally has.

Our final speaker, last but not least is John M. Wozencraft of the Naval Postgraduate School. He graduated from the US Military Academy, West Point in 1946 and joined the Signal Corps of the US Army where he stayed until 1955. He then went to MIT in the EE Department and stayed there till 1977, and he's currently at the Naval Post Grad School working in areas of command and control Today he'll be telling us about a time-slotted algorithm for packet radio networks.

JOHN WOZENCRAFT This work was done in conjunction with a thesis student, Carl Tritchler, a captain in the Marine Corps. He did a really outstanding job of programming to get the simulation going. Let me explain quickly to you what the problem and some of our assumptions were See Figure 1. I'm indebted to the preceding speakers who have covered a lot of relevant material that I won't have to spend time on. First, our view of the problem is that the military want to talk. I'm not convinced that they need to, except maybe psychologically, but certainly they want to. Therefore, the approach we have taken into looking at spread spectrum networking is that there's going to be a voice requirement even at the very forward echelons, and that the military probably isn't going to be able to afford two different communication systems, so that it seems worth a certain amount of effort on our part to see if one can accommodate voice communication within a packet radio spread spectrum network format. When you do that the traffic volumes which we believe are fairly representative indicate that the bits per second for voice are going to absolutely dominate the bits per second of data, so we've just forgotten about data for the time being and are looking only at the voice problem. To do that we're assuming time-division multiplexing on the transmissions using parameters which seem to be typical for vocoded or CVSD voice, a millisecond per slot using about 12 slots per frame. Each conversation goes on of course for many seconds, so each frame you've got to come back and send another millisecond devoted to a particular conversation, and within that millisecond you have to compress the data for the whole frame time. We are using a virtual circuit approach. I think we relax that soon to see if it makes much difference but the virtual circuit seems like a natural way to go for a voice dominated traffic anyway We are assuming that blocked calls are going to be lost rather than queued for simplicity. A point which John Olsen just made which I would like to reinforce is that we started this work a few years ago by trying to ask what the propagation problem is really going to be and we are fortunate in having a terrain model of the Fulda Gap in Germany, which is an area that has been studied intensively from a military point of view. This terrain model enabled us, since we had foliage as well as elevations, to make some reasonable estimates of what the propagation losses would be for a reasonable physical deployment of military forces on the ground. And since most of the students involved happened to be Marine students we ended up with a rather anomalous situation of having a Marine Amphibious Brigade deployed in the interior of Germany, but I suppose stranger things have happened.

Anyway, it doesn't really make that much difference. Early on, we tried out a bunch of different deployments, actually 3 different successive positions, unspecified as to whether we were advancing or

withdrawing, but 3 different successive positions. The conclusion which we unfortunately reached was that the whole system faced a very real and a very severe mutual interference problem; i.e., with the kinds of bandwidths which you could reasonably expect to have available, with the kind of deployment and the kind of terrain you could expect, and with the number of people talking you could expect, the channelization of traffic was such that there were some very big transmitters which were going to be jamming a lot of very little transmitters.

Most of the work which we've done since then has been aimed at trying to figure out how one might be able to get away from that problem. Our normalization has been to assume that every transmitter transmits at whatever power level is necessary in order to have a received energy per bit of 1 Joule. So there's very definitely power control here. We've normalized on the receive end instead of the transmit end. Clearly, if you then decide to up your transmit power uniformly, keeping with that convention, say from 1 Joule to 10 Joules, you haven't affected the mutual interference characteristics of the network. You have gained A-J protection by 10 dB and you have lost 10 dB of low probability of intercept protection. Basically, our view is that what you want is to have an operational parameter which says to you for the network as a whole whether A-J is more important or LPI is more important, and adjust your power accordingly. But for all of this, if you don't get rid of the mutual interference, you've cut your margin for both AJ and LPI down.

What we did on this particular piece of work that I want to talk about today was to see if there is some sensible and reasonable approach to getting relatively high traffic density consistent with voice traffic across a simulated network using code division multiple access techniques. As outlined in Figure 2, we are assuming that every transmitter has it's own pseudo-noise sequence which is unique to it, i.e. it's a one-time tape and every sequence is picked randomly from everyone else's and the law of large numbers applies. So unless you've got bad luck, you can in fact receive several different transmissions at the same spot on the ground simultaneously. On the other hand you can't transmit at the same time you are trying to receive, and you can't transmit to more than one intended recipient at a time in the model we are considering. What we are trying to do is to introduce into the routing algorithm, provisions both to adaptively accommodate the congestion within the network, and also to control mutual interference by minimizing the total amount of energy that's transmitted over the network. The (artificial) network we are looking at is not the Fulda gap one, it's a nice little symmetric Test Network, as shown in Figure 3. Typically we are thinking that if the nodes represented military installations on the ground they would be in the order of 2 to 5 kms. apart, quite close together, with the premium being put on transmitting at as low a power level as otherwise makes sense.

There are two basic kinds of problems we have to worry about. One of them has to do with putting the energy minimization into the algorithm, and that has to do with the routing. First however, let me cover the concept and assumptions which we are talking about. I've mentioned a good many of these already. We are assuming that each node can hear and indeed monitors each of its neighbors, so he knows when they are transmitting. He therefore knows that he does not want to transmit to them when they are already transmitting, because they won't be able to hear him. We are assuming that you are monitoring everything all the time

although you only accept traffic which indeed is directed to you, either for delivery or forwarding to another node. And we are making the usual assumptions that the traffic is Poisson distributed, geometric in length, and evenly distributed over the network as far as source destinations are concerned.

The connection algorithm which we are talking about is outlined in Figure 4. You think in terms of a simple little network as shown there, the toy network with 4 nodes. We assume that a call arrives at A and is destined for C and that the routing algorithm says to try to go to B. There are 3 different kinds of service messages that get passed in establishing that connection. R1 tells B what slots are available at A to transmit to B. R2 tells A what slots are available at B to transmit to A, and R3 assigns a slot for transmission from B to A. The algorithm is pretty simple. A tells B in R1, when A can transmit. B doesn't know that, because B doesn't know in which slots A is already receiving. That's why you have to send that information. B picks out the best A to B slot and tells A to use that one in R2, and also tells A which slots B is receiving in. And then A selects what slot he wants B to transmit in and that locks up the A/B (two-way) connection. Having established the link from A to B, B will play the same game trying to establish the B to C link. The thing that is really important to emphasize is that the attempt here is to try to get high traffic density, and therefore we are exploiting in this model the fact that one can in fact receive two or three or four different transmissions from other nodes in any particular slot. Each node will therefore try to assign a transmission slot to his neighbor which will fall on top of traffic that he is already receiving, thereby keeping as many slots clear as possible. The limiting value, k, of the number of signals that are allowed to be received in a single slot is a variable parameter in the simulation. k=1,2,3,4 were the experimental values we tried.

The question of how to do the routing remains. We used a standard minimum distance approach using Dijkstra kind of algorithm, with the calculation done off-line independent of the simulation. How often the min-distance calculation was iterated is again a parameter of the simulation. We were not trying to look at fast ways of doing the calculation, but rather were trying to see whether the general methodology has any merit or not. As indicated in Figure 5, however, there are two things to be considered: One is the link congestion, and the other is the link loss. For our model the link loss (based on our Fulda Gap data) ranged between about 80 dB for the best links and 140 dB for the worst ones. Thus there's about a 60 dB dynamic range in what was called "useable" links to include in the network, which is very much larger than the dynamic range of the link congestion. To combine the two, we proceeded very artificially. We just took 128 bins and assigned the links to bins randomly, according to a finite-geometric probability distribution. Whatever bin a link fell into was taken as the link loss, which was therefore a positive number between 1 and 128.

The link congestion was also done in a very ad hoc way, just coming up with something which would work. The basic idea was to find out how many received slots in fact, would be compatible between the two nodes; if there were a lot of those then the link was uncongested. We wanted to put these two distance measures (loss and congestion) together with a scaling so that when the net was uncongested, we followed low loss paths, and only when the links got very congested would you start using the lossier links. We did that by using a function f as indicated in Figure 6. The

congestion measure was normalized to lie between 1 and 128. For low congestion the distance assigned to the link weight grows at a low rate and therefore the effects of the loss component are more significant. For high congestion there is a steep slope going up to 1024 instead of up to 128, so that the congestion clearly dominates the loss at the high end of the congestion scale. It is the sum of this nonlinear function and the loss which is then the distance assigned to each link, and which then enters into the Dijkstra algorithm to give minimum distance routing.

The real thing we wanted to find out is whether this procedure was very sensitive, whether it worked at all well, whether it seemed like a reasonable thing to investigate further, and I will now show you very rapidly a very large variety of seemingly uninteresting data. Figures 7,8,9 The data is, however, interesting from my point of view primarily because of the fact that it indeed does not show any very great sensitivities to anything. This encourages us to think that schemes of this sort deserve further consideration and further work, although certainly they are a long way from being settled. This shows you some of the parameters which we looked at. We could update the min-distance calculation every 1, 3 or 5 seconds The average call durations were either 2 or 5 or 10 seconds, and we had a number of channels which could be received in any one slot equal to 1,2,3, or 4. Finally, we had 3 different break points on the nonlinear function, which varies the relative weighting between the loss propagation effects and the congestion effects in the routing algorithm. As a control we used a static routing which was always via the shortest number of hops, which is reasonably close to optimum for the situation in which all the traffic densities between every node pair are equal. Figure 7 shows the percentage of offered traffic for which a circuit was established. What seems significant is that the static and dynamic routing numbers don't differ a great deal. Figure 8 shows the average number of virtual circuits which were active at one time. The network is starting to saturate with 10 sec call durations. We are getting only 55 to 75% of the calls through, and are handling about 13 calls on the average at any one time. Things are very much less saturated with 5 sec calls and slot depth 4; you might figure that about 8 or 9 calls the effective throughput attainable with this network over this range of statistics. Again there is not a great deal of sensitivity in the data, which I find encouraging. The small variations between fixed and dynamic routing I think are explicable in terms of the fact that with fixed routing, the average number of hops is smaller With dynamic routing, when you get congested you try to go around. When you try to go around you use up more hops and therefore congest the network worse.

It seems clear that the question of how you do adaptive dynamic routing is intellectually non-trivial. It ought to be making things better, and in some sense it does, but in another sense it makes things worse because it makes the congestion worse. I think the rationale for the dynamic approach is that in fact you don't know what the statistics are. The stationary routing may be preferable if you happen to know what the statistics are, and if you can optimize the routing for them. But if you don't know what they are, or if they keep changing, clearly you can get in a lot of trouble. You've got to adapt because you don't have enough knowledge to do anything else.

The result which makes me happy about our experiment is shown in Figure 9, which indicates that without effecting things very much in the other parameters, the dynamic routing achieved a noticeable

decrease in the average energy required to go across the network (energy being measured in terms of these bin numbers.) So we did succeed in our objective of routing the traffic via the better paths without increasing the congestion, on the average, noticeably in the network. We conclude that schemes of this sort seem promising, although a great deal of additional work is needed. In particular, what is needed is the beginnings of a fundamental understanding of the performance attainable with "optimal" dynamic routing. Without such understanding, it will remain hard to say whether results such as those presented here are really on the right track or not.

FIG. 1

PROBLEM DEFINITION

VOICE COMMUNICATION DOMINANT PROBLEM

DATA SENT IN INTERSTICES

TIME DIVISION MULTIPLEXING

- 1 MSEC PER SLOT
- 12 SLOTS PER FRAME
- EACH CONVERSATION OCCUPIES A PARTICULAR SLOT IN EACH FRAME ON EACH LINK

VIRTUAL CIRCUITS

- PATH AND SLOT ASSIGNMENTS ENDURE UNCHANGED UNTIL END OF CONVERSATION
- UNSUCCESSFUL CALLS ARE LOST, NOT QUEUED

NEED TO CONSTRAIN TOTAL TRANSMITTED POWER, IN ORDER TO

- MINIMIZE INTERCEPTION & DF
- MAXIMIZE JAM RESISTANCE

OBJECTIVES

. TRY AN ADAPTIVE CDMA PACKET RADIO ROUTING ALGORITHM

. EXAMINE EFFECTS OF ENERGY MINIMIZATION

FIG. 2

CONCEPT & ASSUMPTIONS

EACH NODE XMTS USING A DISTINCT PN SEQUENCE (CDMA)

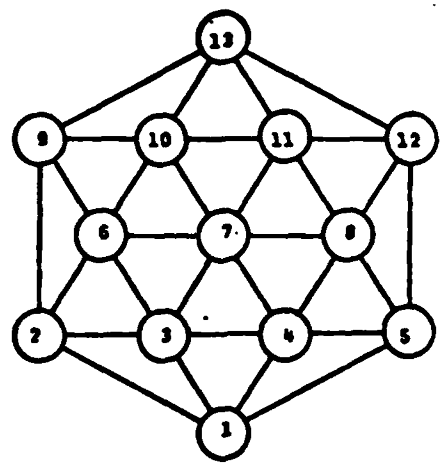
NODES CAN

- XMT ONLY ONCE IN EACH SLOT
- RCV UP TO K TRANSMISSIONS IN EACH SLOT, PLUS ONE SERVICE MESSAGE
- RCV ONLY IF NOT TRANSMITTING IN SAME SLOT

A NODE HEARS EACH OF ITS NEIGHBORS' TRANSMISSIONS, BUT PROCESSES ONLY THOSE ADDRESSED TO IT (FOR DELIVERY OR FORWARDING.)

POISSON TRAFFIC, EVENLY DISTRIBUTED OVER NETWORK

FIGURE 3



Test Network

Fig. 4

CONNECTION ALGORITHM

ASSUME

CALL ARRIVES AT A FOR C

PER ROUTING ALGORITHM, BEST NEIGHBOR IS B

DEFINITIONS

- R1 MSG LISTS ALL A → B SLOTS AVAILABLE AT A
- R2 MSG ASSIGNS A → B SLOT, AND LISTS ALL B → A SLOTS AVAILABLE AT B
- R3 MSG ASSIGNS B → A SLOT

ALGORITHM

- A XMTS R1 TO B IN FIRST MUTUALLY AVAILABLE SLOT
- B SELECTS BEST A → B SLOT, AND XMITS R2 TO A IN FIRST MUTUALLY AVAILABLE SLOT
- A SELECTS BEST B → A SLOT, AND XMTS R3 TO B IN ASSIGNED A → B SLOT

EXAMPLE

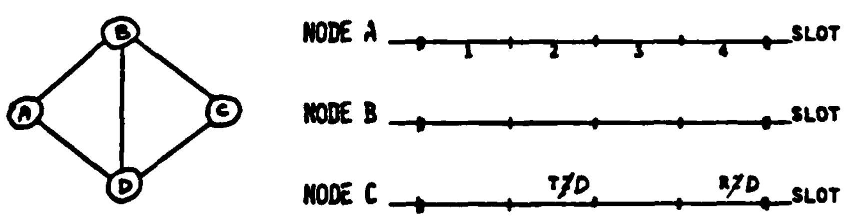


FIG. 5

MINIMUM DISTANCE ROUTING ALGORITHM

LINK LOSS

REPRESENTS ENERGY LOSS OVER LINK.

BINS NUMBERED 1 TO 128

EACH LINK ASSIGNED TO A BIN IN ACCORDANCE WITH A FINITE GEOMETRIC PROBABILITY DISTRIBUTION, WITH STATISTICAL INDEPENDENCE

BIN NUMBER IS TAKEN TO BE LOSS $\triangleq$ L

LINK CONGESTION

X = MAX TOTAL NR OF POSSIBLE RCV SLOT ASSIGNMENTS AT THE TWO ENDS OF LINK

= 2 * NR OF SLOTS * STACKING DEPTH

Y = TOTAL NR OF COMPATIBLE RCV SLOT ASSIGNMENTS AVAILABLE AT THE TWO ENDS OF LINK

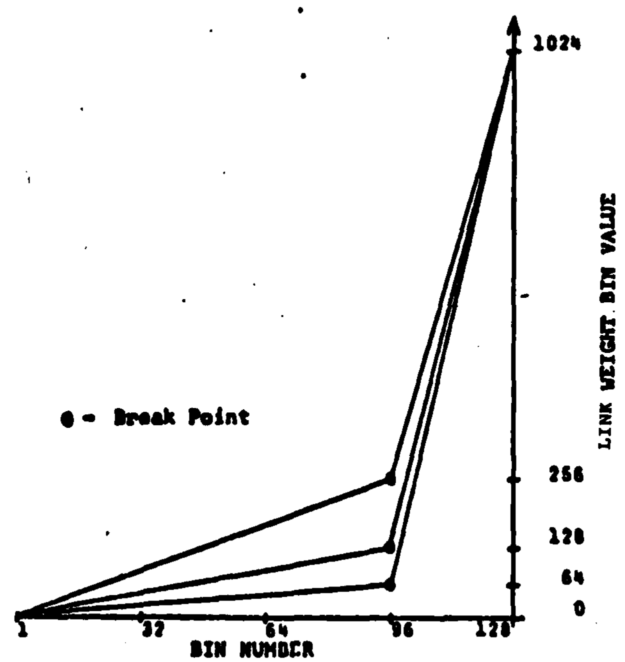
$$Z = \frac{X-Y}{X} \cdot 128$$

LINK CONGESTION $\triangleq$ C = F(Z)

WHERE F IS A NON-LINEAR WEIGHTING

LINK DISTANCE D = L + C

FIG. 6



Link Weight Bin Values and the Break Points

FIG. 7

PERCENTAGE OF CIRCUITS ESTABLISHED

			Average Virtual Circuit Duration								
			2 sec			5 sec			10 sec		
	Update Period		1	3	5	1	3	5	1	3	5
		Slot Depth									
Static Routing		1		97.1			80.9			60.0	
		2		99.6			90.3			70.8	
		3		99.4			94.4			72.8	
		4		99.4			93.2			76.1	
Dynamic Routing	(96, 64)	1	90.5	91.9	88.6	76.1	72.6	71.0	55.8	51.2	52.1
		2	98.3	97.1	95.4	84.7	84.4	81.9	66.6	63.5	63.1
		3	99.0	96.9	95.9	92.9	81.0	86.5	70.8	70.8	66.8
		4	99.0	99.4	97.1	94.6	87.5	84.2	71.0	73.2	70.1
	(96, 128)	1	95.0	93.2	91.3	75.5	75.7	75.5	60.2	56.8	53.3
		2	99.2	97.9	96.7	90.9	88.4	87.3	70.3	71.0	68.1
		3	99.4	09.2	98.1	94.6	91.3	89.2	70.1	73.9	74.5
		4	99.4	98.1	97.5	97.7	92.7	90.9	72.2	71.6	74.5
	(96, 256)	1	97.5	91.9	91.3	81.1	74.9	76.6	55.0	54.8	55.4
		2	99.8	96.1	96.7	94.6	92.7	89.8	73.7	72.2	67.2
		3	98.8	99.2	97.7	95.8	95.0	94.4	76.8	74.3	75.9
		4	99.8	99.4	98.6	97.1	93.8	92.3	79.3	77.0	76.6

FIG. 8

AVERAGE NUMBER OF VIRTUAL CIRCUITS ACTIVE AT ANY ONE TIME

			Average Virtual Circuit Duration								
			2 sec			5 sec			10 sec		
	Update Period		1	3	5	1	3	5	1	3	5
		Slot Depth									
Static Routing		1		3.7			7.8			11.5	
		2		3.8			8.7			13.5	
		3		3.8			9.0			13.9	
		4		3.8			8.9			14.7	
Dynamic Routing	(96, 64)	1	3.4	3.5	3.4	7.3	7.0	6.8	10.6	9.9	10.0
		2	3.7	3.7	3.6	8.2	8.1	7.9	12.7	12.0	12.0
		3	3.8	3.7	3.6	8.9	7.8	8.3	13.3	13.3	12.7
		4	3.8	3.8	3.7	9.1	8.4	8.1	13.5	13.7	13.3
	(96, 128)	1	3.6	3.5	3.4	7.2	7.3	7.2	11.4	10.8	10.2
		2	3.8	3.7	3.6	8.7	8.5	8.4	13.3	13.1	13.0
		3	3.8	3.8	3.7	9.1	8.8	8.6	13.4	14.1	14.3
		4	3.8	3.7	3.7	9.4	8.9	8.7	13.9	13.8	14.2
	(96, 256)	1	3.7	3.4	3.4	7.8	7.2	7.3	10.6	10.4	10.6
		2	3.8	3.6	3.7	9.0	8.9	8.6	14.2	13.7	12.8
		3	3.8	3.8	3.7	9.2	9.1	9.0	15.0	14.1	14.4
		4	3.8	3.8	3.7	9.3	9.0	8.8	15.4	14.9	14.8

FIG. 9

AVERAGE ENERGY FACTOR PER ESTABLISHED VIRTUAL CIRCUIT

		Slot Depth	Average Virtual Circuit Duration: 2 sec, Update Period 1	2 sec, 3	2 sec, 5	5 sec, 1	5 sec, 3	5 sec, 5	10 sec, 1	10 sec, 3	10 sec, 5
Static Routing		1		1.51			1.38			1.37	
		2		1.51			1.54			1.48	
		3		1.51			1.46			1.42	
		4		1.53			1.51			1.55	
Dynamic Routing	(96, 64)	1	0.15	0.21	0.12	0.49	0.60	0.47	1.21	0.84	1.06
		2	0.11	0.15	0.21	0.59	0.53	0.55	1.07	1.10	0.79
		3	0.17	0.19	0.13	0.66	0.75	0.57	1.03	1.04	1.03
		4	0.11	0.13	0.11	0.60	0.63	0.32	0.87	1.13	1.02
	(96, 128)	1	0.46	0.39	0.43	1.16	0.99	0.65	1.09	1.37	1.09
		2	0.33	0.44	0.37	1.07	0.72	0.78	1.27	1.10	1.26
		3	0.29	0.39	0.29	0.76	0.68	0.69	1.26	1.25	1.11
		4	0.22	0.36	0.42	0.74	0.60	0.74	1.07	1.21	0.96
	(96, 256)	1	0.72	0.70	0.63	1.11	1.14	0.97	1.17	1.26	1.03
		2	0.56	0.66	0.61	1.02	1.08	1.07	1.21	1.27	1.31
		3	0.57	0.63	0.72	1.04	0.94	0.88	1.22	1.43	1.19
		4	0.51	0.62	0.79	0.89	0.92	0.98	1.15	1.20	1.15

SPREAD SPECTRUM NETWORKS

DISCUSSION

LEINER: I'll take this opportunity to read the first question. The question is, "Why don't networkers take more seriously the oft heard line that there are three kinds of lies, white lies, damn lies and simulations. My point is that some of the simulations seem awfully artificial." Perhaps the other panelists have a comment. What I'd like to say is, I personally happen to be very skeptical about simulations but I believe there is some value when you are trying to understand some global issues that would change my understanding or trying to get an understanding of what's going on. But to use them to predict the performance of what you might get in the real world is very dangerous.

STEIN: How about relative performance of different concepts.

LEINER: I still think that's the same thing.

STEIN: Still think they are dangerous? You've got a better way to do it?

LEINER: Yes. Sometimes. It really depends. My experience has been that often you can build the system and test it in the field before you can get a simulation that's accurate enough, up and running to reflect the actual performance you'll get.

LEVITT: The problem with the field test is, is it repeatable? Can you duplicate the test and get the same results?

LEINER: And the problem with the simulation is, does it reflect what the situation you actually see in the field is? I think you need both is the real answer. But I think relying on simulations is very dangerous.

STEIN: For absolute values, but if people are going to advocate different concepts, and the only other kind of evaluation is going to be gut feel because you have nothing to back it up, I think relative performance in simulations that people can agree have some of the right kinds of features, it ought to be a lot better than nothing.

OLSEN: That's true but often you can do a combination of generic modeling and analysis that gives you a better understanding of what's going on than an analysis alone would.

STEIN: I'm not sure I believe that.

OLSEN: I'd like to comment on that. I don't think that pure analysis could come up with predicting any kind of performance of a complicated network in a dynamic environment. It's just too complex a problem to do an analysis and simulations, I just can't think of anything other that can give you a first cut feel of are you going down the right approach. I can't go along with going out and coming up with the design and going into manufacturing of a system without having a quantitative feel for how it's going to perform. But whether the simulation you make your initial decisions on, whether that reflects <u>exactly</u> what the performance in the field is, I doubt it very much. The problem we traditionally are confronted with I brought up during my talk, that's the propagation modeling which is such a major factor in predicting performance. The propagation modeling could easily be off by a factor of 2 in terms of how quickly the links are changing or how many links exist. But one of the things we can do is feed back results from a field test into our simulations to try to correct the modeling inaccuracies in terms of

what propagation we experience in the field. We are seeing in our field tests that our propagation models that exist right now are very pessimistic in predicting what an airborne jammer does to a ground based system. This is because the simple models only characterize the terrain, and typically this results in the jammer's free space loss to community, which is very pessimistic in terms of the jammer's effectiveness in jamming the community. What we've seen in our field tests is that in fact the terrain does quite a bit more to help for the ground community in masking the community from an air-borne jammer.

WOZENCRAFT: I would hate to make all the mistakes I've made in the simulations in real hardware first. I think when we go to make the hardware we'll make enough additional mistakes then. My view is that you are certainly going to need both. Wherever you can put field data into the simulations you are obviously way ahead.

GRAFF: I have a question here about one of my viewgraphs. It says, "Here in Air Land Battle 2000 someone mentioned dynamic net connectivity, and your talk described network." That should really mean dynamic network connectivity. So the network connectivity is changing not the net is changing. Does this mean that the Army is willing to support the notion of independent orgar.ic nets to getting all their communications via a common user network." I don't know how to answer that question. Needless to say, the Army does need to define need-lines for Air Land Battle 2000. The current need lines are need lines which have been defined over the past and that's what the current systems are being built towards for fielding. What I described was a technical concept of a potential spread spectrum network for the ALB 2000 environment. Additionally, the network connectivity may change, but the net (i.e. the people who typically communicate) may <u>also</u> change with time and as the battle progresses.

LEINER: The notion of internetting and having a common user network doesn't mean that the particular units have to subordinate their requirements. They can still have their own organic communication. What's important is that those networks can be internetted with other networks so that in addition to supporting their own organic location requirements they can support the requirement to communicate with units outside their units.

LEINER: "In a jamming environment how can you sense the channel in order to determine if the given channel is or isn't in use?" Larry, who were you addressing that to?

LARRY: Anyone.

WOZENCRAFT: I think that if you are talking about MFSK, you can look at the empty channels and see what the average energy is in them. If the average energy is bigger than some number or is changing fairly rapidly, something is happening, and it's probably jamming or other interference.

LEINER: I think there are two ways. In addition to sensing the channel which I think you can do in some cases for example if you have AGC and you check bit sync and you have some sort of algorithm which sits on top and takes a look at the state of the world as it's being perceived, perhaps you can tell whether the difference between interference, jamming and noise and just now signal. But in addition there is the question of whether you need to sense whether the channel is busy or not at all. For example if you use an ALOHA hop protocol, you don't sense whether the channel is busy, you just transmit. The way you find out whether you are successful is if you hear an acknowledgement later on. So there's

a case where you don't bother sensing the channel at all. In a sense that is more "robust".

MILSTEIN: If you do have a scheme where you need to sense how do you do it. That is to say, how accurately can you do it if indeed you're being jammed as well as if in the process of sensing you sense the jamming as well as a possible other user using the channel you're anticipating using.

LEINER: Oh, in other words, how can you tell the difference between a true user and the channel being used. That you can do because you can tell the difference between energy and bit sync. That combination will allow you to tell the difference.

OLSEN: I would tend to disagree with that. I would think any information that you require for net management algorithms to know whether the channel is being used by your system or not is infeasible in terms of being too highly subjectible to spoofing. I don't think CSMA is appropriate at all to use in an anti-jam network.

LEINER: Are you saying that you can't develop an algorithm that's non-spoofable that can tell whether there's jammer or not that's an interesting point?

OLSEN: Assuming, as I stated that the intelligent jammer knows everything about your system other than your keys, he can pretty well mimic your waveform exactly.

LEINER: Suppose you have to have the right waveform in order to get bit sync. For example in packet radio you have to have the right code to get bit sync.

OLSEN No. He's still going to lock you up until you determine that he's not the real system.

LEINER: No, you don't, it just never comes through the matched filter. If the matched filter just sits and listens for the code that is right for that time slot, and the time slot is short, ... Actually we should move this discussion off-line.

PURSLEY: One problem of that is that he can simply regenerate, retransmit that code if you are not changing it fast enough, he can relay it around the network. It's just a replica of your own code that you've transmitted. I think it's a real problem with CSMA in direct sequence.

LEINER That's actually why we chose 5 milliseconds.

PURSLEY: Yes, but it actually has to do with the geometries of the jammer and the communication links and whether or not he can simply delay and then relay that back to you and lock up the network with your own code, your own signal.

WOZENCRAFT: I don't see any reason not to consider in principle at least situations in which you have one-time keys. If we are talking research for 10 years from now, my guess is that's going to be economically feasible. In this case I don't see there's a problem.

PURSLEY: I don't understand, how does one-time key get you out of it?

WOZENCRAFT: Each node has its own key stream, its own pseudo-noise sequence, and that pseudo-noise stream never gets reused.

PURSLEY: If we have multiple terminals out here who wish to have capabilities of transmitting to one particular receiver, then how would you do that? <u>Is he going to have multiple receiver that has been transmitter oriented?</u>

WOZENCRAFT: The received station is going to need a copy of the key of every transmitter that he wants to receive

from.

PURSLEY: For every possible transmitter?

WOZENCRAFT: For every possible transmitter that he wants to receive from.

LEINER: Question for me. "What is your code validity interval?" The answer to that is it's downline loadable over the air is the concept. I'm not sure how to answer that question. That depends on what we finally end up setting our algorithms to. "How do you achieve initial synchronization?" We have algorithms for doing network time synchronization that were developed in the context of another radio that we developed. I'm not sure that they will work all that well. I think that area needs a lot of work. Not just the initial synchronization but how do you maintain a network in close enough time synchronization so that you can do this time-slotted operation that I described all through the network. The notion is you achieve initial synchronization by basically, it's a matched filter, and you can just sweep back in time. The other thing is we have time-stamping on the packets so that allows us the mechanism for synchronization.

QUESTION TO PANEL: "Does voice store and forward have potential for military networks?"

LEINER: Yes.

OLSEN: Of course, but voice is a very high capacity user and it is very difficult to support on a large scale in a network. It definitely can be supported. If you have 1000 users and they all want voice, they cannot all be supported simultaneously, you can support some degree of voice communications.

GRAFF: I think 16 kilobits CVSD voice, certainly 2.4 LPC is a potential.

OLSEN: If you're talking about spread spectrum, of course you are trying to keep the data rate down. Whenever you significantly increase the communications requirement in terms of baseband bandwidth, it's going to cost you in spread spectrum improvement.

Even 1200 bits per second voice is still a large requirement on a network because you have many users trying to share the channel simultaneously. Voice has a limited application but certainly I don't think you'll see a radio spread spectrum communication network that will provide voice interconnection of all users in a battlefield.

LEINER: I can see an application in overlaying if you have a data network and you need to support a limited amount of voice on it. We've demonstrated that to do LPC or even 16 kilobits CVSD for us but as you say, it's going to be a very limited number of effective conversations.

GRAFF: It's certainly no problem to accommodate into the network.

STEIN: I'd like to comment on that because I've heard the voice, or no voice going on for so long. Obviously in the tactical world there are two kinds of voice. There's the kind of voice that the tactical air force talks about where delays are intolerable. There's the other kind of voice which you can really question whether it's needed except maybe among commanders. And if I can tell a tiny story on that. About two years ago I participated in a civil aviation study, looking at replacing air traffic control over the ocean by a data link, and all sorts of yells went up from the civil air community who loved their voice single sideband. Sitting around in the panel and discussing this there were a number of people who had been around 20 years ago when they tried to introduce the voice single sideband to replace the teletypes that were then on the planes and they remembered the yowls that went up

then. That was their cynical answer to how bad voice was really needed.

WOZENCRAFT: I think that there is some basic human need for voice. I think that a commander is going to want to be able to talk to a subordinate in trouble, get a feel for how bad the situation is, and I think the subordinate in trouble is going to have a psychological requirement to share his problems with somebody. I'm not prepared to say whether the military "needs" voice or not, I'm only commenting that they are human beings, and people have been talking for a very long time. I think there are a lot of times where data is very much preferable, and where the military will think it's preferable; fire control, for example. But there's going to be a residuum of circumstances in my opinion where the military benefits of voice justify the additional problems.

GRAFF: I'll just add a few words to that. The Army has been rather voice-oriented obviously in the past, but the Army is building quite a few battlefield automated systems which have very strong obvious requirements for data, real-time data. And that's what's driving a lot of our current thinking. We also see that trend increasing rather than decreasing over the next 20 years or so, so I cannot really comment on whether the Army will or will not use more or less voice. That's not a developer's prerogative but I can say the Army is building more and more automated data systems.

REED: Why do you eliminate the possibility of two-way voice communications establishing two-way voice communications, simultaneously? Is there some reason for that? I got the impression from these various talks that you only establish one-way links in voice.

GRAFF: No, that was the network connectivity I was addressing in terms of an adaptive AJ environment where you can go one-way with a certain kind of characteristics, but the reverse path has different characteristics in terms of its jamming environment, how you use throughput, that kind of thing. You can create a two-way logical path from one-way physical paths, provided that the appropriate links are available.

REED: But ordinarily you establish two-way links?

GRAFF: Yes, most times.

STEIN: User requirements are generally two-way because of the end-to-end protocols if nothing else, but the link level connectivity may very well be one way because of the jamming and other different environmental considerations.

REED: But in the usual instance it would be two-way, like an ordinary telephone exchange system?

LEINER: One other observation on this, it seems to me there is a fair amount of work going on both in overlaying data systems on voice communication systems and in overlaying voice on data. I think what's going to happen as we evolve is one of two things. Either the two of those are going to become so close that it doesn't matter or there will be a better refinement of the requirements so that we understand what the data rates that have to be supported for voice versus data are and that will govern which approach is the one. But I certainly agree that we are going to have to have an integrated communication system as indicated by some of the other speakers.

STEIN: I think you should also not overlook the very rapid progress that is taking place in automated voice entry systems which will really reduce the data rates. At ICASP 2 weeks ago one of the hits of one of the sessions was a game which is now going on the market in which the game participant actually speaks

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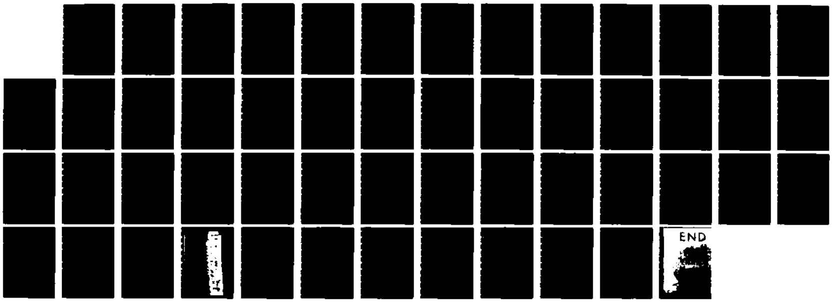

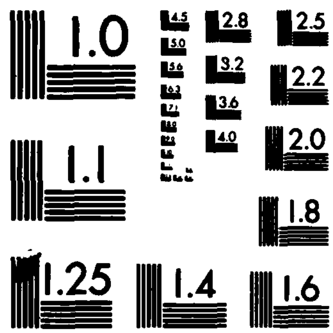

MICROCOPY RESOLUTION TEST CHART
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in a limited vocabulary after a very short training session.

LEINER: Another question. "Is CSMA the only good candidate for packet radio?" The answer to that is: No, it is not a good candidate for packet radio. The reason is when you move into the bit-by-bit code changing environment where people are transmitting on different codes it's impossible to sense the channel because he may come up in the middle of a packet and try to sense the channel, you'll discover that it's "not busy" except that it is really busy, but in middle of the packet. "Does token ring have any potential?" I don't think so because I believe that token ring requires you to know who to hand off the token to and in the kind of tactical mobile environment we have to work in, we generally don't have a very good knowledge of who your neighbors are

WOZENCRAFT: I'd like to disagree to the extent that in order to send service messages to establish circuits, you've got to be able to find a channel on which you can transmit to your neighbor and he can receive you. You can't do that if he's transmitting in the same slot, so you've got to be able to sense whether he is or not. If you want to call this CSMA then I think that it will be needed, and I don't see any reason why it should not be available.

LEINER: Normally when people talk about CSMA, I think they mean real time sensing of the channel and transmitting instantaneously based on whether the channel is busy or not. In that respect, I agree with you that a protocol for establishing if you are going to divide the overall channel into subchannels and establishing when channels are busy and who should be able to use channels, certainly you are going to have to have a protocol to do that.

"With regards to low-cost packet radio, is VLSI implementation of various network functions being considered." We have VLSI-ed FEC as you heard from Joe Odenwalder. The real tradeoff there is the initial startup, of course, and so far we haven't build enough low-cost packet radios to justify that. We are only building 1000.

CAFARELLA: There is really another problem in that too, in that you can apply VLSI to a lot of hardware functions which may become more and more sophisticated. But typically the network processes remain in software not simply because that's how it's easiest to initially implement them, but they always change. Almost continually, to the point where you probably would never really want to lock them into a VLSI chip unless it was basically another microprocessor, something that is still inherently very programmable.

LEINER: Let me say that there have been people who have asked the question of whether TCP and IP, now that they are DoD standards should be implemented in VLSI. If we wind up with a very stable TCP and IP, that may be worth doing.

TOBAGI: I was interested in asking this question because also it has some implication as to the time it takes to process packets through PR which happens to be a limiting factor when it was microprocessor based.

LEINER: I think that's when processors were slower in the past. Now they are much faster. The 8086 that we are using in low-cost packet radio really isn't. We don't think we'll slow up the channel.

GRAFF: In certain types of network protocols like X.25 there's a commercial type chip available Western Digital 2511 which does quite a bit of the X.25 protocols for you. It's a DMA controller and does most the link level handshaking, connection management and auto retransmissions. So it's not unreasonable

to put some of the network functions in VLSI eventually.

LEINER: Actually we are using the 2511 chip in a low-cost packet radio because of its tremendous advantage in amount of hardware to do a link level protocol, to the HDLC protocol.

Another question for me. "If the capacity of a packet radio network is not sufficient to support the user requirements what are the system parameters that can be changed not necessarily dynamically in order to increase the network capacity? In other words data rates, topology?"

I think the first thing we can do to dramatically change the capacity of the packet radio network is to have multiple networks working in the same area. The first thing that comes to mind, to me, is working on different spreading codes, but in fact it is much easier to just operate different frequency nets in a given area. For that reason the low-cost packet radio has software selectable frequency on a quasi-packet per packet basis. It takes about 7 milliseconds to change frequency. So it's actually feasible to run a packet radio, a single node in two or more frequency nets simultaneously although you are not going to have full access in any one of them and you may lose some packets, but end-to-end protocols can take care of that. Data rates, yes we can increase the data rates dramatically. We are doing a lot of work with MIT Lincoln Labs, John Cafarella, to build high bandwidth radios, high bandwidth in the RF bandwidth sense and high bandwidths in the data rate sense. Topology, that gets into things like power control. If you can control power of the radios so that they have the minimum connectivity required in order to keep a fully connected network, a network that is totally connected, not every node directly connected to every other node, but at least the capability of routing through the network, you can get the spatial reuse frequencies of codes spatial reuse of the channel throughout the network and therefore increase the capacity that way. But I think the first, the most easy, obvious and trivial thing to do is simply operate multiple networks in different frequencies.

WOZENCRAFT: I guess I've got to disagree again. That might be the easiest and quickest thing to do, but I can't believe you aren't forfeiting performance by doing that over taking the same amount of bandwidth and making it all available to the entire network and putting in appropriate control algorithms. I grant that we may not know what the control algorithms are yet, but in principle you are bound to be better off maintaining homogeneity as opposed to partitioning the network into sub parts.

LEINER: I absolutely agree. It's simply a question of hardware feasibility. For example, the low-cost packet radio can be tuned over 140 Megahertz. We are developing radios that can operate at 140 MHz RF bandwidth. The latter case is certainly going to be more efficient if we can control it. The problem is that it's also more expensive to build the radios.

REED: Could I ask another question? I was thinking about what Seymour Stein said yesterday about using directional antennas to possibly help with the jamming and interference problems. This network reminds me of a neural network. A neural network is a network very much like the network you're talking about except it's not a broadcast network. Each node does not broadcast omni-directionally, each node broadcasts in a directional manner. I wondered what you had done to think of possibly using directional capability, where since you are at L-band for switching directions when you get to an interference problem. And you can use the fact that you can establish geometry by using transponder

codes and that kind of thing to find out where the distance to your neighbors are, positions and so forth. Have you considered that possibility or generality.

LEINER: I'm glad you asked that question. We have taken a look at building adaptive antennas to work with packet radios, and as I mentioned yesterday, had a great deal of difficulty with that. It's not that we don't believe in the concept, I think the concept is a good one.

REED: ....packet radio, to be able to communicate in this direction, and switch to this direction. Point beams in different directions.

LEINER: I think there is a potential problem there in the tactical environment, because just because there is a unit in one direction from you physically doesn't mean that the direction in which the propagation path has to go or should go. Perhaps you can argue that if you are using the channel itself to determine the position, then it's apparent position is in the right place and it could all work out. I think that's worthy of exploration. We have another program that's called the multiple satellite system, is our current name for it. What it basically is is packet radio in the sky. The idea is to build a survivable proliferated satellite system by launching up 100s and 100s of low-cost satellites. (Laughter) You laugh ... it turns out to be cheaper than one high-cost satellite. And the idea is to use packet radio-like protocols to do store-and-forward, so the ground user accesses the system, it gets relayed over to the proper ground entry point and then comes back down again. In that system, where line of sight is clearly defined, there is no problem with multipath in space. We are planning, actually, on using adaptive arrays, programmable arrays, to actually steer the beams in the direction of the intended recipient, because it gains us a lot in range for low-cost. So that is a good thing to do. The question, is there some way of controlling it. If you're moving around in the ground and your connectivity is changing very rapidly and not in a predictable manner, as it would be in space, is it possible to do that kind of steering of the antenna beam.

OLSEN: One point is just a practical aspect of beam forming in a tactical environment. There are certain users that are fixed locations that could afford to have an array antenna, but the majority of users that are highly mobile and really couldn't afford to have an array, at least physically to carry it, certainly a man-pack could not carry an array.

REED: Well, you are talking about a L-band and so you could have a horn that you could rotate actually, you could consider rotation of a horn.

OLSEN: Well, I certainly wouldn't think that you would want to have someone ......

REED: You could get something like 10 or 15 dB gain.

OLSEN: There's no question you can easily get the gain, but the pointing of it, if it's not electronic, I don't think it would be appropriate for a packet switched network where you are changing antenna direction 180 degrees from millisecond to millisecond. The tactical manpack users can't carry them around, the weight is already at a premium just to carry around the radio as it is as opposed to carrying multiple antenna elements or a mechanically steered antenna.

STEIN: Maybe Cafarella can comment on it because I am not familiar with it, but the paper I was referencing was actually talking about an electronic steerable microwave antenna, not a very heavy system, something that could be put up on a mast. Do you know about it? It's at Lincoln somewhere, I don't know what

they were developing it for.

CAFARELLA: I think what you are talking about is the antenna that was developed for Camp Sentinel radar in Vietnam. We later built a ground surveillance radar using that antenna. Fortunately the antenna is about $50,000 and rides on a truck, so I think you have a problem deploying those kinds of antennas. Originally the concept was cheap but when it was finally built it was not cheap. So it could be a problem. Maybe if you built it, if someone else built it ....

STEIN: The one I was talking about, they were talking about a microwave version. Camp Sentinel was way down in frequency.

CAFARELLA: It was later built, I think the advanced ground-surveillance radar was at C-band. So it was built, the elements were made using printed circuit board technology. The individual vanes were made of printed circuit board on glass epoxy, and the whole thing was stacked together so that it was a really big cylindrical thing. Had another problem, that the way you steered the beam was by switching, and there's lots of loss in switched matrix too, so if you look at it, for the extra power he's got to put out, you'll lose a little bit in LPI too. But even with the printed circuit technology for fabricating the antenna, it came out pretty expensive.

GRAFF: I think we are ready to acknowledge that a distributed ground environment doesn't mandate omni directional antennas. I think we are trying to right now assess the impact on AJ, LPI, multi-user in a network environment.

STEIN: If I can comment on that, one reason for looking at the directional antennas is every time you look at any of the threats that people put up, you find that what they are talking about over large numbers of jammers, distributed in every which direction. And the directional antenna would obviously help on that score.

LEINER: If there are no other questions, let me ask a question of the audience. This is a new culture, we'll have the panel ask the questions of the audience now. The panelists I've assembled here are representative of work that has generally tended to be, if not tactical, at least ground and air mobile. The question of spread spectrum networks also comes up in a satellite context though. Would anybody in the audience comment on whether combining spread spectrum with a satellite system gives you more or less, or equal to the straight addition of those two?

OMURA: I guess in some sense they do use a little of CDMA capability to some degree as I understand it. There's hopping of groups of users and then within the group also you may have several users that are in a sense using multiple-access and so there's kind of two levels of scrambling and mixing. One is hopping very wide to avoid AJ, the other is hopping within a limited band to have CDMA basically. That is sort of the idea behind the systems.

LEINER: Do the same or similar systems that we run into in managing that kind of environment, that spread spectrum environment with different codes and different keys and all, come up in the satellite systems?

OMURA: Well, you don't have the complicated propagation conditions. I thing the propagation conditions work, really makes it so much harder on the ground.

LEINER: The fact that it's store-and-forward repeater, incomplete, moving topology.

OMURA: Your network is also just

one location up here, going up to that or cross linking, but the number of terminals it is just very simple compared to what you have on the ground. Propagation-wise, the number of terminals and network configuration, nothing dynamic.

LEINER: So when we go to things like MILSTAR, or this multiple satellite system that we are talking about, we are liable to run into similar types of issues.

RISTENBATT: Except that MILSTAR is a small number of very expensive objects, synchronous, and the multi-satellite situation is more like packet radio where all of the things we are worrying about come up. You have a very dynamic situation because you have relatively short orbits, periods for the orbits, and the connectivity is changing all over the place, and also the burden is on the satellite network itself to organize most of the routing and traffic management whereas in MILSTAR it's all ground control. In the synchronous systems you still put all the control down at some place, or a few places that are sanctuary and do all the control from there.

LEINER: So are you saying that the real critical issue here is the survivability issue and the requirement for distributed management?

RISTENBATT: I guess all I'm saying is I'm impressed with the problem of trying to control the chaotic situation associated with packet radio or multi-satellite situation, whereas the conceptual control of a satellite system with a few objects in synchronous orbit is got to be much simpler. We do it.

I'm nurturing an idea which may be hare-brained. But you ask, what kind of possible synergism is there between a satellite and a spread spectrum, and I'm thinking something like this, couldn't one make something of the bus sort of architecture of the presence of the satellite. Here's how spread spectrum could, for example help. That broad bandwidth, you could user multipath to your favor, for example, you could literally use pulse position for signaling, you could sync to whom you wanted to and if the satellite were good enough. I don't know if that's superior to code division multiple access, but it seems like a satellite and spread spectrum have something to offer, you know the wide bandwidth. It seems like that there should be something there. Has anyone pursued anything like this?. I would imagine that something like this has been done.

LEINER: Are you suggesting doing code division multiple access by not necessarily code division, but maybe the same code with pulse position and use the resolution you can get with spread spectrum to be able to refine the different components and demodulate on the chosen one?

RISTENBATT: It's the resolution I am paying attention to. I am wondering if you wouldn't get a great relief from any kind of centralized network management until a certain point, the system, you have a broad time bandwidth, you are talking about that's the way to use this, is to fill up. You don't have to direct a lot of people. I really don't know how this would work, but if you have a.... I know there have been various attempts over the years to show that if you really wanted to multiplex the number of people, who aren't on continuous or bursty on something to fly, there's no way to organize to make that efficient except to let it run free and use the whole TW space. And it's always been hard to grab hold of what they've actually shown. It's difficult stuff, but I'm just wondering if a satellite and spread spectrum doesn't offer the promise of a pretty much of a free running network and you get the most out of it. When it gets overloaded, it gets overloaded, because no

system can defy gravity. But I'm just wondering if there isn't something there

LEINER: Thank you all for coming.

SESSION 5 - INDIVIDUAL PRESENTATIONS AND WRAP-UP SESSION

WEBER: I thought that during this final session we would get a few view points from some of the participants about what they thought of this workshop, whether we should do it again in the future, should some part of it be classified,..., so I've asked a few of you to comment. In addition, if anyone else would like to comment, we do encourage that.

I decided it would be best to start with the person whose idea it was to have this workshop in the first place, namely, Bill Sander from the Army Research Office; so Bill, if you care to come forward and make a few comments.

SANDER: The first thing I want to do is express my personal appreciation for everybody who has taken the time to come out and participate in the workshop, and that's particularly true for those panel members who prepared and gave presentations, and even more especially for Bob and Chuck who worked so hard over the last year or so putting this workshop together. I don't know how you feel about this workshop but I feel very good about it. I think we've had an excellent workshop, and particularly, I think the facilities and the environment have been outstanding. I'd like us to give both of them a hand this morning for all of what they've done during this past year. (Applause.)

I do feel, as I said before, that it's been a good workshop and I think there will be a good proceedings coming out of it. When we started this thing, I was thinking more in terms of spread spectrum, particularly more in terms of specific details of coding and different performance comparisons, and in what directions my program should possibly take. But I'm not disappointed in what's happened here. One thing that has really struck me is that we have a lot of knowledge about specific techniques, narrowly focused techniques, and not enough on how these techniques fit together in the whole system, and that seems to be much of what the discussion has been about. I think that is something that we need to consider a great deal and focus on in the future.

Let me change gears a little now and say something about the Army. There have been some questions asked about different aspects of utilization of spread spectrum in communication networks in the Army. ALB 2000 has been mentioned. ALB 2000 means that people whose jobs it is to predict these things, are predicting by the year 2000, that if we were to be involved in a major altercation, then it would be a very rapidly moving battlefield situation, different from anything we've experienced in the past, not a guerrilla warfare type thing in Vietnam, not a slow plotting type war as we had in World War II (for a major war), but very fast moving. What this means to the planners is that information must move fast to the commanders in order to enable them to make their decisions in near-real time so as to react to what's happening in their battlefield. That's why we have the requirement for the kind of high throughput, robust, communication networks that we've been discussing during the last couple of days. Now we proceed on from here, and we go on to the next step: How do we achieve this? Well, thinking in a very general level, there have been two approaches mentioned. One is simplicity: Make the network and the equipment as simple as possible, the approach as simple as possible, so we have reliability and keep the cost very low. But this doesn't consider the second problem of ECCM and mutual interference and those are very very real problems. And not only that, it

doesn't consider the fact that GIs today can't operate the World War II equipment they have in the field. So what most people believe is that we must go to much more complex approaches to solving these problems. Reliability and cost are big questions, but we do need the ECCM and we need automated operation of these networks, automated as much as we can so we keep the GI out of having to twiddle the knobs, since if we let him twiddle them, he's going to twiddle them the wrong way. That's typically what we find out. In the long term, I think the Army communication systems are going to spread spectrum, packet radio, and distributed communication systems. By distributed I mean distributed control, distributed routing, distributed management, distributed for survivability and with adaptivity involved. The major question is how to choose from among the design alternatives. Do we know enough about performance in general to make design decisions? What I've been hearing here in the discussion, if we know enough about it, there are a lot of differences of opinion. We don't really have a common agreement on how to go about doing this. Partly I guess it's the Army's fault for not better identifying the threat conditions that the system has to satisfy, and the environment in which it has to work. Maybe this has to be done on a system by system basis and not in a general sense. We have problems defining these requirements in a very realistic way for the systems that we are concerned with for development.

In conclusion, I want to summarize the way I think the research opportunities are heading, after having listened to the discussions over the past 2 days. One is in adaptivity: Antennas, power control, code changing, whatever ideas you can come up with in regards to automatically adapting the network to the varying conditions in the threats that it's going to be subject to in an Army environment. Another is acquisition and synchronization; I think this is an underlying problem in any spread spectrum system and it is a very realistic one, and the more we improve that, the more potential for higher performance in the networks. Whole system design; I've already mentioned that. I feel like we need to consider the big picture, and not closet ourselves in little areas of technology without maintaining sight of the whole picture. Try to keep abreast, as much as you can, about the big picture. Coding is mature but there is still a lot of work that can impact a lot of areas, such as multiple access, maybe even synchronization and acquisition. Simulation was a question that was also discussed and I personally believe that we need simulation, but we need that simulation to be very realistic, even to the point of taking real data and replaying it through the simulation rather than trying to come up with artificial representations of what we think the situation is going to be. We need to make the simulation as real and accurate as we can. I don't see how we can experiment with all of the design alternatives that we may want to consider; therefore I think simulation is going to be essential. We have simulation in computer-aided design of integrated circuits, we now have simulation in imaging with image processing, we can simulate model images. In my view, there is no reason why we shouldn't come up with some of these same approaches to simulating and modeling the communications that we are all working in. And finally I think something which is not relevant to the group here, namely, there are a lot of technology issues which, if solved, would make the job a lot easier. In particular signal processing, VLSI chips that could make the job easier, wideband amplifiers, power amplifiers, wideband antennas. It is very realistic that we need much more research support in: Accurate

clocks, fast frequency sources, etc. You can probably think of more than I can. Thank you Chuck.

WEBER: Thanks Bill. I realize a couple of you have to leave early; and I'd like to get your comments. Barney, would you please make a few comments.

REIFFEN: When I was invited to participate in this conference, my first judgment was somewhat skeptical because I thought that the really substantive issues of today in spread spectrum systems really couldn't be well confronted in an unclassified conference, but having participated in these few days of meetings and listened to what everybody said, I think I'm going to change my mind. I think the idea of this workshop is very good. It really represented an opportunity for different points of view, research people, academic people, some people involved in system implementation to share views and just cross-pollinate and that's really what the purpose of a workshop ought to be. I think that some of the really substantive issues in AJ system design were in fact joined, particularly in the third and fourth sessions, and it was very well done. That's one comment I wanted to make.

The second comment, perhaps a little controversial, is that it's my judgment that the themes of modulation/demodulation theory and coding/decoding theory as applied to spread spectrum systems is really pretty well understood by many people. Perhaps not all, but many. That doesn't mean to say that there shouldn't be any further work in this area. Certainly there will be, but I would not expect the work to give rise to anything fundamental. I think one can expect only marginal improvements in what can be done. I think the thing that paces system performance at this point is not a lack of theory but its use in implementation and cost and matters of that sort. In fact, when one undertakes to design an AJ system, there's an awful lot of attention focused on the modulation and coding part of it but the real tough part of a job usually turns out to be the problem of time acquisition, frequency acquisition, angle acquisition if we have to point antennas. That represents probably the intellectual challenge of designing a well-conceived antijam system. It is sort of self-obvious that when you design such a system, the acquisition must operate at or below the threshold SNR that you expect to communicate in. That's a test that has to be applied to one another of the acquisition schemes. In that context, I found myself a little perplexed when there were some discussions of adaptive systems that assumed that the signal had been acquired in order for the adaptation to take place, where in turn the adaptation was necessary in order to operate at the threshold SNR. So there's a little bit of a dilemma there, and unless I'm missing something, I think there is a little bit of chicken and egg in that thinking.

One other comment, I think that the workshop at its best does represent an interchange between theoretical people, analytical people, and people in charge with building and realizing systems. I think this workshop was fortunate to have had the opportunity to hear Col. Gobein, who gave to us, a fascinating case study of a real system in the real world. The politics of it, the changing requirements of it, the competition among different approaches through it.

I think I would make one specific recommendation about any future conferences, meetings or workshops on this theme; namely, that an attempt be made to do a case study in depth, even further depths than Col. Gobien did so that the person leading the discussion can sort of say this is the problem we tried to solve, this is how we solved it, these were the tradeoffs that were made, that choice

was good or bad, and it brings some real world factors into some of the theoretical calculations, and in that spirit I think I'd like to echo what the first commentator said.

WEBER: John Wozencraft, I know you need to also leave early.

WOZENCRAFT: A few comments. The first one is to again to congratulate the organizers and sponsors of the workshop. I think it's been certainly one of the most rewarding professional experiences of this sort that I've encountered. I'm pretty tired of big massive conventions where there are so many people there that you rarely have a chance to get into any real discussion in depth. I mean, that with a small group like this in isolation, the opportunity to explore, both during the question and answer period, and also around the swimming pool, is very rewarding and very much appreciated. It's particularly fine for someone like myself to have a chance to renew acquaintances with colleagues and co-workers from earlier decades and to make the acquaintance of a new crop of perspective colleagues and co-workers; a great opportunity, and very much appreciated.

Coming now to the more substantive issue: Should we have additional workshops of this sort? I would certainly come down with a resounding "yes". I think my own preference would be for a classified session, maybe next time, perhaps alternating between classified and unclassified workshops. Possibly the timing of about once every couple of years would be my best feeling as to how fast progress would be made so that it's efficient to have a meeting like this again.

To make a quick comment about the subject on simulation which came up again. I am obviously very much in favor of simulation. I think that many of these problems are entirely too complicated to come up with closed form analytic solutions or even open form analytic solutions. I even believe in making realistic simulations but I don't believe in making only realistic simulations. When you go to treat a problem analytically you make as many simplified assumptions as you can in order to get the "know" of what it is that you want to study. And I think the same thing is absolutely essential in making progress in a simulation world if you're trying to cover all the water front at once before you have pinpointed and identified your problems and your concepts and your perspective solutions. You'll be absolutely awash with the sea of data which even a statistician would have trouble making any sense out of. So I think you need a whole wide range of simulation activities, just as you need a whole wide range of analytical activities to support this kind of venture

Finally I will be rash enough to spend a minute or two trying to put together, or express my own internal synthesis of the sense of this workshop. It seems to me that the laws of physics, as well as the presentations which we had, come to a fairly clear conclusion that one single transmission communications network is probably not going to suffice for military tactical communications. You have the understandable feeling on the part of the infantryman that he would like to be very close to the ground, and he would like to be behind cover and hidden by foliage, and the laws of physics say that electromagnetic waves at high frequencies and high bandwidths don't like to propagate under those conditions. So it seems to me that you are going to need a fairly low frequency, low UHF/VHF, maybe going all the way down to HF or MF depending on the adversity of the conditions. With very mobile equipment, very small and light-weight equipment, and therefore not very capable equipment. This sub-part of the system is going to have to be very mobile indeed The trouble

with that is that if you want to get the advantages of the strong connectivity of networks, and their really wide-ranging abilities to reconnect in a robust way in the face of adversity, I don't think you can get that from here, because these are going to very low-capacity sub-networks. I therefore envision higher frequency, higher capacity sub-networks; let's call it a backbone network. (At least that's what we've called it in the work we've done at Naval Postgraduate School (NPS).) This backbone sub-network would be at L-band, or C-band, or somewhere up higher which uses the same ideas of packet switching and routing, and which interfaces, if you will, through an internet. But I don't really view it as a gateway, just a change of transmission facility using the same formats, etc. everywhere. For these higher capacity nodes which are perforce going to be less mobile than the very mobile backpack type of equipment; I'm talking here about jeep mounted type of equipment. And here, it seems to me, is where you have the legitimate and important need for spatial processing which you can't get in the size and mobility requirements of the backpack guide. But here you can in fact have 10 or maybe even 20 dB antennas, you can have adaptive arrays that can null out enemy countermeasures as well as some of your own mutual interference problems. These guys would move around; they'd put their antennas, (which would be at high enough frequencies so that they're fairly small), up above the canopy of the trees. The very mobile guys would attach themselves adaptively and dynamically to the nearest of these less mobile backbone stations. And there would be the facility that, if you couldn't get there in *one* hop, then you would packet switch from small transportable station to small transportable station until you could get into the backbone. Once in the backbone, the traffic would be all put together, and you'd then have much higher capacity shipping it around, until again then you have to leave the backbone in order to get back into your ultimate mobile terminals. I don't really know that this is going to be economical. I am of the opinion that if we are going to be economical, it's because people *learn* how to really build much more complicated stuff a lot more cheaply than we know how to today. I think that these problems involve a high degree of logical complexity in order to gain the operational simplicity which was talked about as being required. But really, I believe we've got to make things that are so terribly complicated, that in fact, they are simple and transparent from the point of view of the user. And the equipment pretty much takes care of itself in terms of adaptivity.

I would disagree with Barney Reiffen a little bit or at least add one more comment to his, reiterating what I said yesterday. I think there remains one very severe intellectual barrier to being able to build and understand and analyze the kind of system which we've been talking about. And that is the understanding of the adaptive dynamic network control in the face of adversity. I think you've got to have the acquisition but I view that as a link level problem primarily. You've got a big network problem on top of all that. I, at least, am not going to feel comfortable until I see an understanding of that which exceeds any that I've encountered so far.

__WEBER:__ While the mike is there, Seymour, do you want to add something.

__STEIN:__ Yes, I'll try to resist the temptation to say anything technical because I already had my shot, and talk only about logistics and workshops. First I would like to put in a plea, particularly to the Army if they can do anything to subsidize a rapid conclusion to the proceedings and get it disseminated before the next workshop takes place.

(Laughter) I'm looking forward to getting a copy and I just envision that the people at USC have taken on a real bear of a job with getting people to proofread what they think they've said, etc. I really believe the proceedings will be of value to everybody and I'm looking forward to getting them.

Secondly, on the workshop itself, I have to echo what everybody else has been saying. I think it's been a great experience. I too have the feeling it ought to continue and I think there are some very real questions about how you do continue a workshop type environment, which is what we are all asking. This is especially true when you think of the hordes of people who attended the MILCOM convention, also on the topic of spread spectrum, and how many more of them will come to the next workshop when the word gets around that we've had a great time. (Laughter...) Frankly, there always is a problem with workshops. How do you keep the size down by keeping the mailing list very secret? (**WEBER:** It was hard!) I know. We could very easily double the number of people here, all who would have been able to contribute and derive quite a bit from this. Along those lines I have a few practical suggestions. First, that the original theme of this workshop be mentioned, which as I heard it, was looking for future research directions, not the near future, but the long term. Certainly that limits the scope of the kinds of people who really ought to attend. Secondly, I think another way to limit attendance, (this is, I think, what Jack (Wozencraft) is perhaps suggesting) and that is picking narrower topics, one or two topics per workshop, rather than an across the board type of thing. I think this is well on its way to becoming a once every year kind of thing because there are certainly enough topics. I also agree that at least some of the sessions, or some of the workshops perhaps need to be classified. I think all of us have a feeling, and I think it's a legitimate feeling that jammer design, interceptor design, and communications design are all really interrelated. They are all best done by people who really understand what the other side is doing. When you try to design an AJ system, it's always very illuminating to realize that the guy designing the jammer can't do all the things you think he might be able to do because he has lots and lots of problems too. At the same time, there's at least one area that I've been exposed to where I suspect there are some interesting lessons to be learned for spread spectrum design and that is the area of detection of signals. Quite a bit of work is going on in this area, all classified, but knowing how we publish things in this country, I don't think it will be more than a couple of years before it's quite diffused in the literature and will also become a legitimate topic for a Ph.D. thesis. I suspect it would be very useful, in one of the early future workshops to make that one of the strong topics; that, I know, would have to be classified. I think I'll hold my comments to that.

WEBER: Thank you Seymour. In terms of getting the proceedings out, it is going to be a task; the turn around time from you people, however, is also a big parameter in that equation. Bob and I have already been talking about how to best expedite this. We'll play with it for a while, and see how fast we can get a first version at it back to you. Ed Posner, you want to comment?

POSNER: This is a very valuable workshop to me and I've learned a lot. It would have been good if we could have had one classified session. I don't think you can have a classified session in this room, so if you could find a military base within 20 miles (Laughter...). (**SCHOLTZ:** I could look for a classified ranch!) (Laughter...) Of course the Army might

want us to hold the whole workshop in Fort Huachuca. That may be a way of holding the attendance down. (Laughter...) Solve both problems at once, get the hard core interested people. But I feel that a classified session of serious communications researchers would have been a good add-on for half a day i we could have found a way of doing that. It might take some creativity. Another advantage might be to have a kind of joint services sponsorship. This has been very valuable for tactical communications, but if there were a way for the Secretary of Defense or someone to sponsor more of a Joint Services meeting with some inter-service sessions and look at some of the broader and maybe some strategic issues in slightly wider community, I think that would be a good service for the government. Maybe the Army could still sponsor it, or run it but if there were some joint participation, I think the benefit to the entire US defense would be perhaps greater and also for the research community. But I'll again repeat that I've learned a whole lot and have a lot to think about and people to work with. I think every year is probably is a good time frame to have it rather than every two years. And if USC can get out the proceedings in 6 months, that would be good.

WEBER: Ed?

BEDROSIAN: When I was young and bursting with ideas, nobody ever asked me. The old goats did all the talking. Now I'm being asked, and I don't know if I like it. My technical comments are not major ones. I'll save whatever I have to say for the later discussion. I would like to comment on the workshop itself. When Chuck asked me to come, I started thinking, reviewing in my mind, how it is that meetings are held, what we try to achieve at meetings. I don't have to tell you. It depends on what sort of result you're trying to present to what sort of a group. If you've done research that's complete you want to archive it while you put it in a journal, and if it's fairly complete and you want to get it out to a fair number of people quickly, you go to a technical meeting. The workshop is different. These are supposed to be preliminary results, frequently tentative ideas, even hare-brained ones if you will, the idea being to expose them to coworkers who can benefit from getting some informal thinking quickly. The workshop does that admirably Many workshops have been tried: most have failed. The reasons are manifold. I don't know what happened here but this one worked. As I look back and ask myself why, one of the best ways I can answer the question is to compare it to what took place last week in Florida, the IEEE Communication Theory group had a workshop there. I've gone to those annually. Seymour was there as some others here were there too. One of the things that I noticed was the attendance. It fell off exponentially with time. The last morning of that workshop, I didn't want to go, but I stuck my head in just to see what was going on and I had to go in out of embarrassment. There were 5 speakers and 2 listeners. I just couldn't let that be, so I had to sit there. I didn't want to (Laughter...) Why did that happen? One of the reasons was that there were 5 or even 6, but every session of 3 hours had enough speakers to fill it completely and the poor session chairman, Gaylord was one who faced this problem, (or did you give a paper .... laughter!!!) Well, he wasn't the culprit, but the poor session chairman would have to stop questioning. He'd say, well, we must move along, we are running out of time, and here are people raising their hands wanting to ask questions. Well, that's not a workshop, I don't know what it is, but here you very successfully kept the talking part by the

speakers down to a satisfactory time. First of all they were good speakers, so we had interesting things to talk about; and there was plenty of time to talk. And that worked. If you could repeat this, well, marvelous. We've got to have more of them on this structure.

The other thing, when I think about what you're trying to do in a workshop, the main thing is to provide the proper atmosphere as well as the structure of the thing. And the purpose of course is to promote personal interchange between people. And now, structurally what you do, you choose the session lengths, the topics, you form panel discussions, you encourage the impromptu presentations and so on. Sometimes it isn't as though there's a cookbook that tells you how to do this. Sometimes it just plain happens.

But there's another part to it that I want to emphasize, and that is that this personal contact that we try to encourage in a workshop ought to go beyond the professional part of it. I notice, as I look forward to these meetings, that it's as much the opportunity to renew the social acquaintances that makes them desirable. In that regard, being able to include family is very important. Maybe others don't share my view but it's just wonderful to be able to talk to people with whom you work and correspond and have these very esoteric discussions on a human level because this is the way we deal with most people in our lives and I think that helps make all of us seem more accessible and we understand one another better. So things like having communal dining, getting away from places where there aren't distractions is great. The cost however, has got to be within reason. And if I'd known what it was going to cost, I didn't find out what my wife would really have to pay till we got here. So the fact of the matter is, the subsidization of the participants, at least from my point of view, nice is almost meaningless to me because my company is going to pick it up anyway. If you could subsidize my family! (Laughter...)

As far as classified sessions are concerned I've been involved in holding any number of meetings, and classified ones are a royal pain in the neck. The things you have to go through to find the place and then to get the clearances, and you have to have security people present and all sorts of terrible things happen. In fact, you have to give up many of the things that makes a workshop successful in order that it be classified. So personally if at all possible, I'd prefer to avoid unclassified meetings, not that I don't think classified ones are desirable, but because of what they do to you. And as far as the written output is concerned, my notion of what the workshop does is centered, as you've got the picture of, around the personal contacts and while I'll be happy to get a written output, personally I doubt very much if I'm going to study it. I'd use it the same way I would journal articles and technical reports that are published. In fact some people have perhaps, realistically pointed out how they are doubtful whether they are going to get anything out of it in a period of time. The truth of the matter is, you might be able to submit these papers to journals and get them printed, in not much less time. That's just the reflection of the practical problems of doing that.

I can't resist one technical problem. That's about simulations. We all barely have our thoughts on them at the risk of arguing with the Army. The idea of making simulations realistic can often be a terrible trap. The one that comes to my mind, perhaps some of you are familiar with it, is the bomber penetration model that the air force has. And it is a blow by blow accounting of what a manned bomber, penetration of the Soviet Union

would be like. I forget what the time interval is, I think it is every minute, maybe it's less. Every minute, they look at every bomber, every air defense site, every interceptor, everything. And you could not be more realistic than that thing is, but it takes hours to run on the fastest machines and it costs so much that you end up running so few cases, that you're not sure what you have when you get through. So doing what Jack (Wozencraft) was talking about before, you've got to simplify simulations down just the same way you simplify down analysis. You've got to get results. And about tactical communications, my own recollections from hiding behind rocks and foliage as Jack was talking about, make me remember something that we as technical people may forget about. That is that when people around you are so angry at you that they are shooting, things get terribly confused. Discipline on almost every level just falls apart, and that system that works very well in the lab or maybe pretty well; when you go out on the field, it will just absolutely fall apart. If you've got frightened people running around, they don't know where they are, they don't know what they're doing, and if your system works, then it's a miracle. I would absolutely always opt for a simple system that will work, for a complicated one that has a better performance. My final comment is, many of you don't know it, but the Germans had individual microwave communications in World War I. They had the antennas, they just didn't get the electronics. It was on the helmet, if you've ever seen a World War (Laughter ...) ... a half-wave dipole.

WEBER: I would just like to thank our USC student helpers, Mary Anne Kiefer and Peter Pawlowski, and Milly Montenegro. I think they deserve a hand for contributing so well to the success of this workshop. (Applause) And actually, the choice of the place here was easy because we knew about it from the Communication Theory Workshop last year, and Bob suggested that either we go here or we don't do it. I agreed with that from the start. Anne Giles, I also think has done a remarkable piece of work here, and I am certain that her efforts make it easier for us to do our job. And I think all of you did too. The cooperation of the participants is what really makes the workshop a success, as we look for things that Ed was talking about that make it all play. Is there anyone else that would like to make a comment?

SANDER: Some questions have been raised about the workshop and how it comes to be and its tri-service sponsorship. This is not difficult. We do this. Each office, probably AFOSR, ARO and ONR, each support half a dozen workshops of this type every year, maybe even more. The way you go about doing this is get in touch with me if you want to organize the workshop say next year, or maybe two years from now, and talk it over with me first, and then submit an unsolicited proposal for doing that. The proposal need not be very lengthy, as Bob will tell you. The preparation of the proposal is miniscule. It's the work that comes after that. If I have the funds and decide to support it, that is the really hard work. So before you send in a proposal or call me to discuss the possibility of organizing another one of these workshops, think hard about how much work it is to do because it's not easy, and then get in touch with me. As far as joint service sponsorship, if you send me the proposal, I can send it to the Navy and the Air Force, (ONR and AFOSR) and if they want to participate in the funding, that's a very very simple matter, that's just a nipper, and is done all the time with workshops also. We receive money from them to co-sponsor workshops and they receive money from us to co-sponsor workshops. I think this just happened to be

the first one on spread spectrum and I just decided to take it all. If ONR or AFOSR had come in and said, hey, we'd like to share in the funding, I surely would have accepted that kind of participation. So that's how it's done. I think the hardest task would be for someone to come up with an idea for having a classified workshop or even a workshop with a classified session. We don't sponsor anything classified, but I think it would be possible to write a contract with someone to have a classified workshop. I would have to check into that. I can't guarantee it. I guess that's the status. If anyone has any questions as to how to do that, I'd feel free to answer them. And I would like to hear from you from someone who would like to have another one of these workshops.

RICE: I found that I wanted to agree with a couple of things, and I want to point out a couple of things that I think were not said, and I kept expecting to hear. For one thing when we talked about jammers, I mentioned this to several people around the pool or so, I expected to hear something about the problem of having a receiver saturated by a jammer. There was some talk about using, that in processing instead of nulling the jammer. In such a case you have to null out the jammer. You just can't handle the problem with processing. Another thing I expected to hear something about is optics, optical processing where you can get tremendous processing speeds. I heard 100 mHz, 120 mHz, those are huge numbers, particularly in the absence of optical processing.

In a classified session, we could talk about some other things. There would be a different emphasis I think on some of the problems. But let me just say that if you are in a situation of having to process some of these signals off-line, you're not going to have much hope with 100 or 120 mHz because there's no way to record it and get it off line, except possibly optically; that is, to do something optically.

Let me suggest also that if we think about a classified session, there are places where you could hold classified sessions that are not very conducive to the kind of interpersonal relationships and social activities that were just mentioned. And I think they are important also, and I think maybe a place like the Naval Post Graduate school or maybe even some university with classified facilities where there's still an opportunity to get together and do things socially; there might be appropriate places.

And I want to mention the way I use simulations. That's simply to tell me when I have a bad idea. If I have some signal processing idea that I want to test out and I simulate a doppler, or multipath, or additive noise, or multiplicative noise or timing jitter, or phase noise or something, even if I haven't done it too realistically, if my idea won't work under those circumstances, I'm pretty certain it won't work in real life. If it does work under those circumstances I'm still not sure it will work in real life, but I feel a little better about it. But I think simulations are valuable from that standpoint, and I can sort of get to that by varying the amount of the doppler or the amount of the multipath or the amount of the timing jitter, the carrier drift or whatever. I can get an idea about how pathological these effects become before my signal processing technique fails.

WEBER: Does anyone else have comments? If not, John Bailey and Irv Reed indicated their willingness to discuss the topics somewhat further, of Monday morning.

GENERALIZED ADAPTIVE NODE CONCEPT

JOHN BAILEY: I am going to make my presentation quite short. 10 or 15 minutes, given though I have enough material for a major session. You may call this impromptu paper #1, spatial adaptivity revisited.

First, just a couple of comments. This is a viewgraph I gave Monday. The point that was being made here is that there is the potential for coping with very wide bandwidth jamming only under the premise that one goes to digital techniques where you can implicitly form and invert the sample covariance matrix. This in turn implies a significant amount of hardware if one is going to solve this type of system of linear equations. What we've been working on in the last couple of years, is a concept that would lend itself to being effected in a digital realization with a single digital circuit that could accommodate a wide class of adaptive problems, including radar, sonar, wideband communications, and adaptive signal processing. The idea being that if you could enable such a single circuit, perhaps in VHSIC, to become viable in terms of cost, it ultimately could be put on even wideband communication systems which is the most onerous of all the systems we are looking at because of their wide bandwidth. So that when I talked about architectures of this type yesterday, where potentially one had to have a real-time processor to do this type of operation, the key issue is whether this can be done with viable hardware. Well, before I get into that, let me jump to the crucial question first. i.e. how viable is this for wideband communications? For cheap systems such as UH, voice, things at L band like the global positioning system, IFF perhaps even datalinks like the MICNS and JTIDS, it's probably not applicable because we are talking about very cheap systems, and inherently just the cost of things like A-D converters, much less something that inverts the covariance matrix, is not a viable add-on to those types of systems. However, there are a class of more expensive communications systems where I see this happening possibly even in the near term. Examples might be the large antennas carried on jeeps that were alluded to earlier and certainly some of the satellite communications systems where you're protecting an uplink or a downlink or a relay for satellites which are inherently more expensive to begin with. As to whether there is an inherent limitation in terms of laws of physics, the current limit right now is not digital circuits to do the matrix inversion but the D-A converter. Currently we can, and are in the process of designing and building systems with up at 50 mHz bandwidth for radar types of applications. We do not see anything immediate in the 100 or 150 mHz regimens simply because the A-D converters do not exist that can cope with those types of bandwidths. That may change in the future. So with those caveats I'm going to very briefly describe a concept wherein one can build architectures and develop a single circuit to handle a wide class of adaptive problems. The advantages of the approach I'm going to be talking about is that a wide class of problems can be handled with a cystolic architecture of identical circuits. What I'm talking about is a single digital circuit that can be wired together in a feed forward manner, with a pipeline architecture. This is very important because that implies that as the dimensionality of the adaptation increases, the throughput rates remain just as fast, just as within the pipeline FFT. It's ideal for VHSIC because if you can solve many of the world's adaptive problems with a single circuit, limited only by the dynamic range and bandwidth of that circuit, then

you have something that potentially services enough customers that it becomes a candidate for VHSIC. What we want to do is to implicitly or explicitly effectively solve a set of linear equations in a very general manner. Having said that, I'll go back to the original simple example I gave yesterday of a system of one adaptive degree of freedom to cope with one interference source. Slide 3 Here is the covariance matrix, this is its inverse. If you multiply it by a steering vector of 1 and 0 you end up with a weighting vector and if you then apply it to the data, you end up in effect doing nothing more than finding the depicted scalar weight. Therefore if I have a circuit that just performs the multiplication, it implicitly does the inversion, determination of the adaptive weight and the forming the ultimate beam output.

Now, in Slide 3 the position of the lines is a notational artifice. When the arrow comes down under a node, what it really means is the steering vector is 1 and 0. The circuit is ignorant. What that circuit does is to take the ratio of the cross-correlation between the two jammers and normalize it by the power in the auxiliary channel. Now the crucially important thing is the orthogonality property. By orthogonality we mean that the adapted output is orthogonal to the auxiliary channel, i.e. if you take the cross-correlation average with the complex conjugate of the auxiliary channel, it's zero, assuming Gaussian statistics. The reason that is important is because if I now form this type of structure using that notation, it means that all of the outputs in Slide 4 are mutually orthogonal, which can be shown by extending the argument for the two channel case. Note that any transformation (T) for which the first row of the inverse is equal to desired steering vector is suitable. And that in turn can be effected via a trivial transformation. This completely generalizes the concept of sidelobe canceller. Now let's see how this Gram-Schmidt approach compares with other algorithms. What I have here is a list of all the algorithms I know of that fit into the class of very rapid convergence because they either explicitly or implicitly invert the covariance matrix that have actually been used in hardware systems, viz. radar, sonar or communications. Table 7 shows the number of computations per unit time that have to be done at the bandwidth of the system, as well as the total number of computations that have to be done over a time basis of 2N samples. Note that the generaled GS technique is faster than all the others. However, the really important point is the pipeline architecture aspect of it, because if one wishes now to increase the dimensionality of this device, you just add more circuits. Thus it is feed-forward and is a pipeline. Having said that, we very briefly describe a fan of its other properties.

If you look at this network and *regard* E *as an input* vector and the output of such stages as an output, then you can show that this really is implicitly a product of simple transformations. What you really have is an upper triangular matrix in a factored form, and that's important because the factorization in this case looks exactly like this where the coefficients are the weights that are actually formed in the nodes. And what that means in turn is that it is also trivial to invert this because the inverse of each factor is the same factor with the off diagonal terms replaced by their negative. As such the inverse is trivial.

So far I've shown a generalized sidelobe canceller. Now consider a system in which you have an array of elements, that generate multiple beams. Figures 8-10 show the substantial savings that is realized by operating in beam space.

Figure 11 shows the Widrow type configuration suitable for utilizing a pilot

signal in communication systems.

One final word, I want to go back to this because my caveat was that one cannot get beyond 50 mHz using digital techniques for wideband spread spectrum systems. The caveat is, that this limitation does not apply when using frequency hopping as a method of getting spread spectrum. So I'll put up this diagram again. It hypothesizes that you have both frequency hopping and you have some sort of coding at each given frequency whether it's a PN code or a formulation of a Reed-Solomon coding like in JTIDS or any other coding scheme. The point here is that because you can make the network work very rapidly, as long as the time interval between frequency hops is longer than 2N times the number of samples times 1 over the bandwidth, you have the property that you can save data, can solve the problem at low bandwidth, and reprocess the data for each frequency step. So for frequency hopping systems, you can indeed potentially dramatically relax the requisite A-D converter speeds. Of course that means the receiver has to switch frequency with the a priori known sequence of frequency steps. You're obtaining a higher bandwidth simply because, in effect, you're not having to process each frequency step in parallel which premises that you know a priori the order of the frequencies that are being received. This in turn implies that you are already somehow synced. So that finishes my discussion and takes you to Irving Reed who will tell you how to combine the process of adapting with synchronization to overcome this problem which I have introduced. Finally, we'll show you an algorithm that can be implemented with this same generalized orthogonalization discussed here.

EXAMPLES OF NODE UTILIZATION

TYPE	APPLICATION
• Sidelobe Cancellor	Active Systems (RADAR
• Generalized Adaptive Array	Active Systems (RADAR)
• Pilot Signal Utilization (Woodrow)	Communications
• Spatial Mapping	Sonar, Radar, Communications
• Conformal Arrays	Sonar, Radar
• Mainbeam Scattering	• Monostatic Radar • Bistatic Radar
• Adaptive Clutter Filtering	Radar
• Space-Time Adaptation	• Clutter & Jamming Rejection
• DPCA	• Clutter Rejection • SAR Satellite • Satellite
• Feedback Configurations (Applebaum-Howell Loop)	• Convergence Rate/Hardware Tradeoff
• Receiver Equalization	
• Geological Mapping	
• Pattern Recognition	

SLIDE 2

DESIGN ADVANTAGES

A. SINGLE NODE CIRCUIT

1. A SINGLE AUTONOMOUS NODE CAN BE CONFIGURED TO ACCOMMODATE VIRTUALLY ALL ADAPTIVE PROCESSING PROBLEMS.
2. SINGLE CIRCUIT CAN PERFORM FOLLOWING OPERATIONS (IMPLICITLY).

$$\left.\begin{array}{lll} \bullet & \text{SAMPLE MATRIX} & M_N = \sum_{i=1}^{N} \bar{E}_i \bar{E}_i^x \\ \bullet & \text{MATRIX INVERSION} & M_N^{-1} \\ \bullet & \text{WEIGHT VECTOR} & \bar{W} = M_N^{-1} \bar{S} \\ \bullet & \text{ADAPTIVE OUTPUT} & B_A = \bar{E}^x \bar{W} \end{array}\right\} \quad (1)$$

VIRTUALLY ALL ADAPTIVE PROCESSING PROBLEMS CAN BE SPECIFIED BY (1).

B. FEED FORWARD PIPELINE

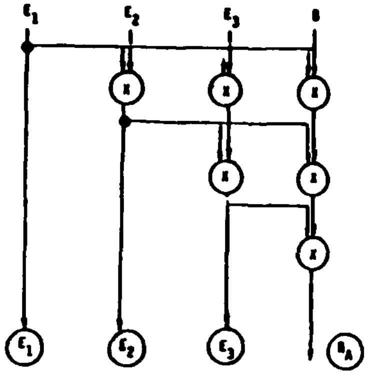
THRU-PUT RATE IS INDEPENDENT OF MATRIX DIMENSIONALITY.

C. COMPUTER-AIDED DESIGN

SIMPLICITY OF MODE AND SYMMETRICAL ARCHITECTURE LENDS ITSELF TO COMPUTER-AIDED WIRING.

D. IDEAL FOR VHSIC

SLIDE 1



IF

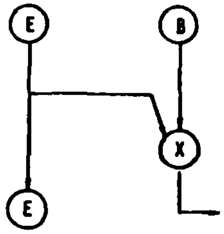
$$M_E^* = \begin{pmatrix} B_A \\ E_1 \\ E_2 \\ E_3 \end{pmatrix} \widehat{\left(B_A^* \; E_1^* \; E_2^* \; E_3^* \right)} = A$$

THEN

$$B_A^* = E^* \left(M_E^{-1} \begin{pmatrix} 1 \\ 0 \\ 0 \\ 0 \end{pmatrix} \right) = E^* \bar{w}$$

GRAM SCHMIDT DECOMPOSITION

SLIDE 4



$$B_A = B + wE$$

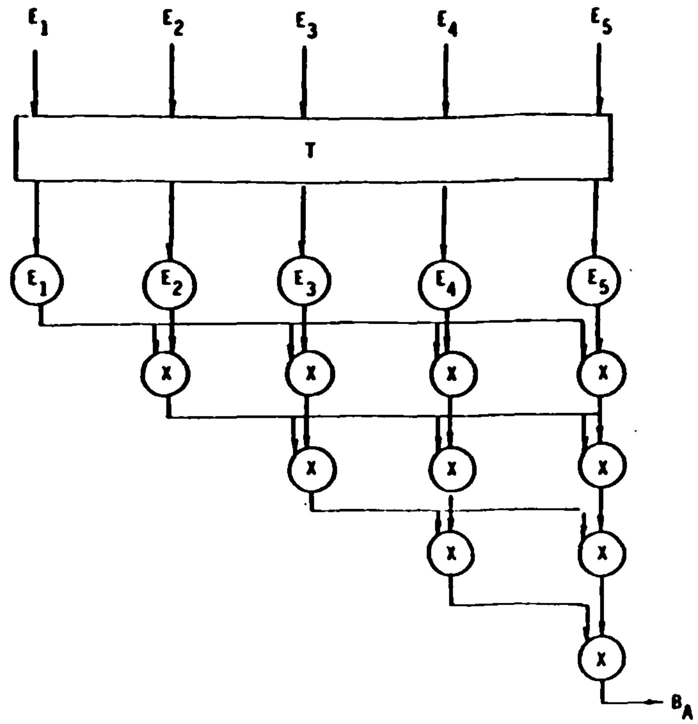
GRAM SCHMIDT ORTHOGONALIZATION

$$w = -\frac{\hat{B} E^*}{|\hat{E}|^2}$$

$$\hat{B}_A E^* = 0$$

SINGLE CHANNEL SIDELOBE CANCELLOR

SLIDE 3



$$B_A^* = \bar{E}^*\left(M_E^{-1}\,\bar{S}\right) = \bar{\textcircled{E}}^*\left[M_{\textcircled{E}}^{-1}\begin{pmatrix}1\\0\\0\\0\\0\end{pmatrix}\right]$$

<u>Any (T) for which</u>

$$\bar{S} = T^{-1}\begin{pmatrix}1\\0\\0\\0\\0\end{pmatrix}$$

<u>GRAM SCHMIDT IMPLEMENTATION FOR</u>
<u>GENERALIZED ADAPTIVE ARRAY</u>

<u>SLIDE 6</u>

$$T^{-1}\begin{pmatrix}1\\0\\0\\0\\0\end{pmatrix} = \bar{S}_N = \begin{pmatrix}1\\ \bar{S}_{N-1}\end{pmatrix}$$

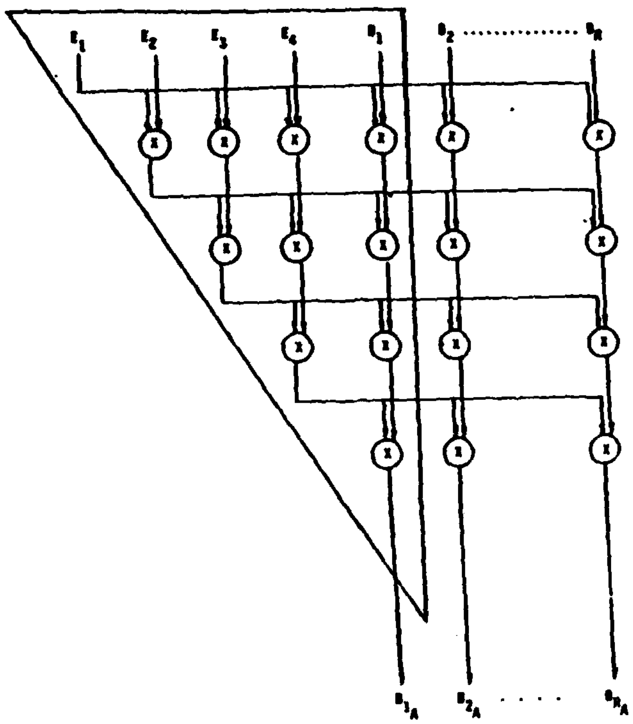
$$T^{-1} = \begin{pmatrix}1 & & & & 0\\ & 1 & & & \\ \bar{S}_{N-1} & & 1 & & \\ & & & 1 & \\ 0 & & & & 1\end{pmatrix}$$

$$T = \begin{pmatrix}1 & & & & 0\\ & 1 & & & \\ -\bar{S}_{N-1} & & 1 & & \\ & & & 1 & \\ 0 & & & & 1\end{pmatrix}$$

- <u>Pretransformation (T) requires only (N-1) multiplications</u>

<u>GENERALIZED CANCELLOR PRE-TRANSFORMATION</u>

<u>SLIDE 5</u>



MULTIPLE BEAM CONFIGURATIONS

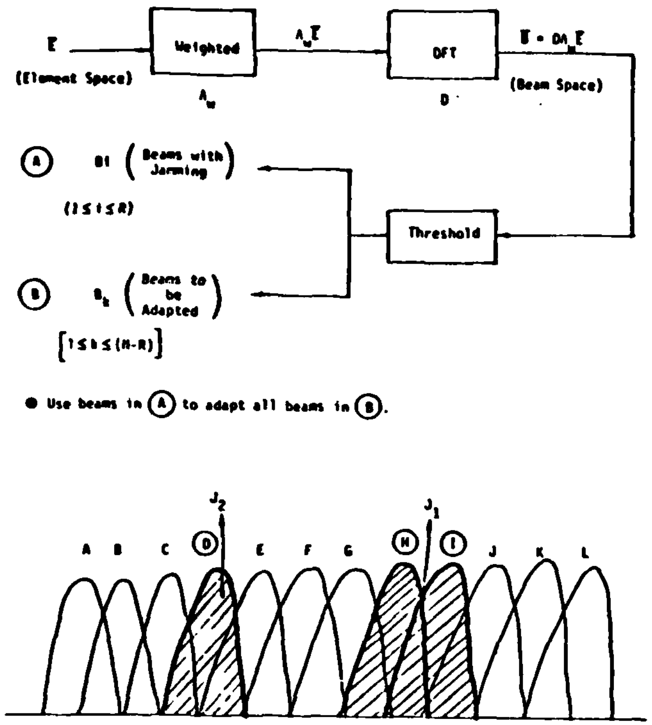
SLIDE 8

COMPUTATIONAL COMPARISONS

ALGORITHM	PER SAMPLE	BATCH PROCESSED (2N SAMPLES)
Sample Matrix Inverse (SMI)	$\frac{1}{2}N^3 + \frac{3}{2}N^2 + \frac{7}{2}N - 1$	$\frac{3}{2}N^3 + 4N^2$
Cholesky (C)	$\frac{1}{6}N^3 + 2N^2 + \frac{11}{6} - 1$	$\frac{7}{6}N^3 + \frac{9}{2}N^2 - \frac{4}{3}N$
Square Root (SR)	$\frac{5}{2}N^2 + \frac{11}{2}N - 1$	$\frac{5}{2}N^3 + 10N^2 - \frac{1}{2}N$
Factored Square Root (FSR)	$2N^2 + 7N - 2$	$\frac{5}{3}N^3 + 9N^2 - \frac{11}{3}N$
Factored Inverse (FI)	$\frac{5}{2}N^2 + \frac{9}{2}N - 1$	$3N^3 + 10N^2 - 2N$
Sample Matrix Inverse Update (SMIU)	$\frac{3}{2}N^2 + \frac{7}{2}N$	$3N^3 + 7N^2$ (NA)
Conjugate Gradient (CG)	$\sim 3N^2$	$\sim 6N^3$
Gram Schmidt (GS)	$N^2 + N - 1$	$\sim N^3 + 2N^2 - 2N$ (NA)

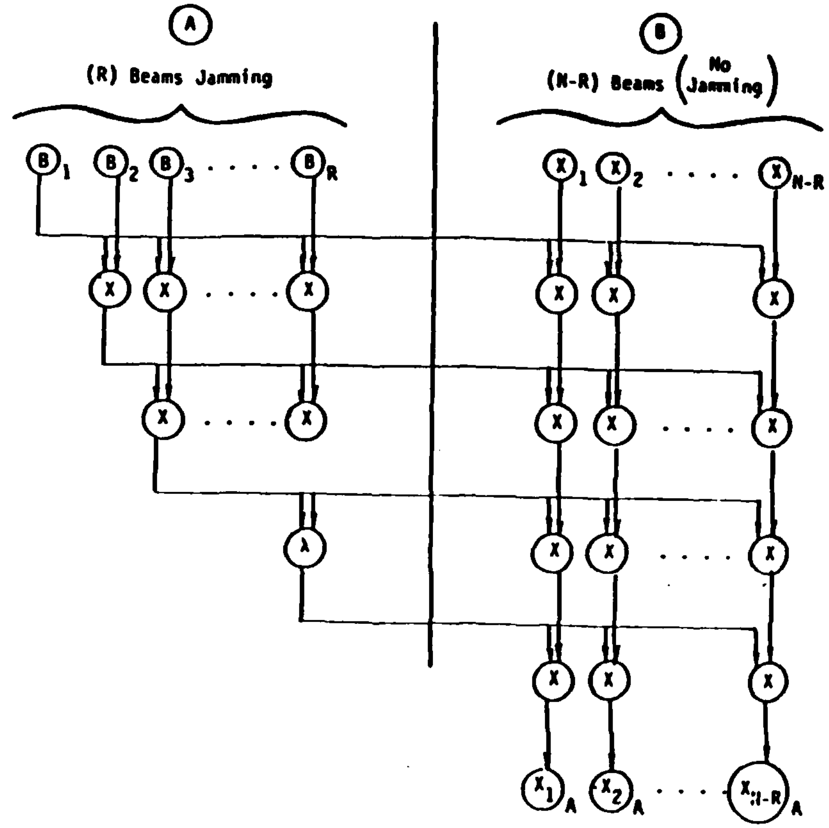
Multiplies + Divides

SLIDE 7



BEAM SPACE ADAPTATION WITH GS

SLIDE 10

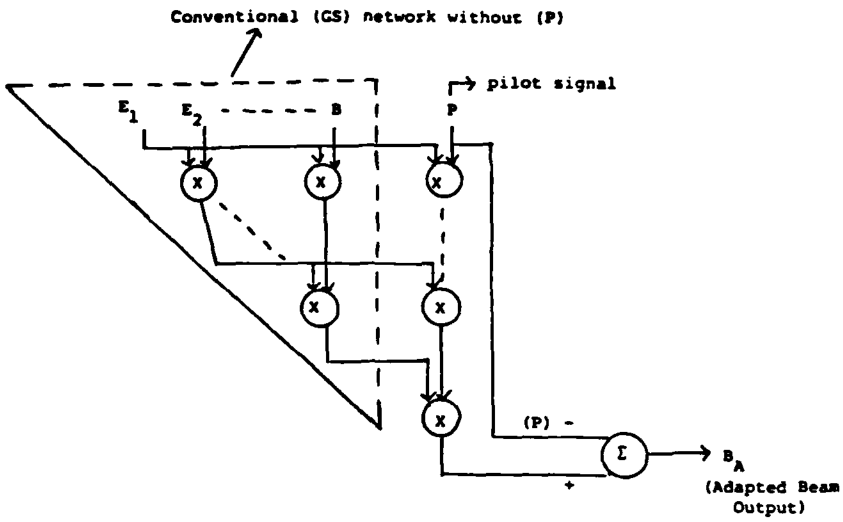


Total # Nodes $= \frac{R(R-1)}{2} + (N-R)R$

Maximum for R=N-1 $\Longrightarrow$ ONE BEAM TO BE ADAPTED

BEAM SPACE DIMENSIONALITY REDUCTION

SLIDE 9



SIGNAL MAINTENANCE WITH A PILOT SIGNAL
(Woodrow Configuration with GS)

SLIDE 11

ANGLE ESTIMATION APPROACHES

(A) • SCANNING

- Null Seeking
- Off Axis Mono Pulse

$(X) = R\left(\frac{\Delta}{\Sigma}\right)$ — Memory → Angle Estimate

(B) • MULTIPLE BEAM

- Multiple Beam Variation of (A)
- Rod Davis Algorithm
  - Amplitude & Phase Distortion of (X) Corrected (MLE)

SLIDE 12

ANGLE ESTIMATION APPROACHES (cont'd)

MATRIX

- **MESA** $M(S) = \dfrac{1}{\left|\bar{S}^{*} M^{-1} \begin{pmatrix} 1 \\ 0 \\ 0 \\ 0 \end{pmatrix}\right|^{2}}$
  - Reciprocal of Adapted Omni Pattern
- **WI** $M(S) = \dfrac{1}{\bar{S}^{*} M^{-2} \bar{S}}$
  - Reciprocal of Average of All Adaptive Omni Patterns
- **ML** $M(S) = \dfrac{1}{\bar{S}^{*} M^{-1} \bar{S}}$
  - $\dfrac{\text{Received Power (Adapted for (S))}}{\text{Unit Signal Power in Direction (S)}}$
- **AAR** $M(S) = \dfrac{\bar{S}^{*} M^{-1} \bar{S}}{\bar{S}^{*} M^{-2} \bar{S}}$
  - $\dfrac{\text{Received Power (Max S/N)}}{\text{Noise Power}}$

- **Eigenvalue** $M = u^{*} \lambda u = \Sigma\, Y_i\, \bar{u}_i\, \bar{u}_i^{<}$

  $\{Y_i\} \rightarrow$ Raid Count

  $\bar{u}_i \rightarrow$ AVG. ϕ Sparial Estimate

SLIDE 13

SPATIAL FILTERING

DIGITAL PROCESSING TECHNIQUES

	PROBLEM	SOLUTION
PRESENT	• Convergence Speed	• Matrix Inverse
	• Clutter & Jamming	• MTI on All Channels
	• Blinking Jammers	• Reprocess Data
	• Wide Bandwidth	• Sub-Banding • Tap Delays • Waveform
	• Receiver Mismatch	• Adaptive Equalization
FUTURE	• Spatial Mapping	• Matrix Alogirithms
	• Mainbeam Scattering o Chaff o Mainbeam Jamming	• Tap Delay Adaptation • Dual Apertures
	• Multipath	• *Sub-Banding* • Dual Aperture • Tap Delay

SLIDE 14

MAPPING ALGORITHMS

MAXIMUM ENTROPY ALGORITHM (MESA)

$$\sigma_1(s) = \frac{1}{|(1000)M^{-1}\bar{S}|^2}$$

MAXIMUM LIKELIHOOD ESTIMATOR

$$\sigma_2(s) = \frac{1}{\bar{S}^* M^{-1} \bar{S}}.$$

<*> denotes the complex transpose operation.

WEIGHTED INVERSE

$$\sigma_3(s) = \frac{1}{\bar{S}^* M^{-2} \bar{S}}$$

AAR ALGORITHM

$$\sigma_4(s) = \frac{\bar{S}^* M^{-1} \bar{S}}{\bar{S}^* M^{-2} \bar{S}}$$

SLIDE 16

COVARIANCE SPATIAL MAPPING

ADVANTAGES

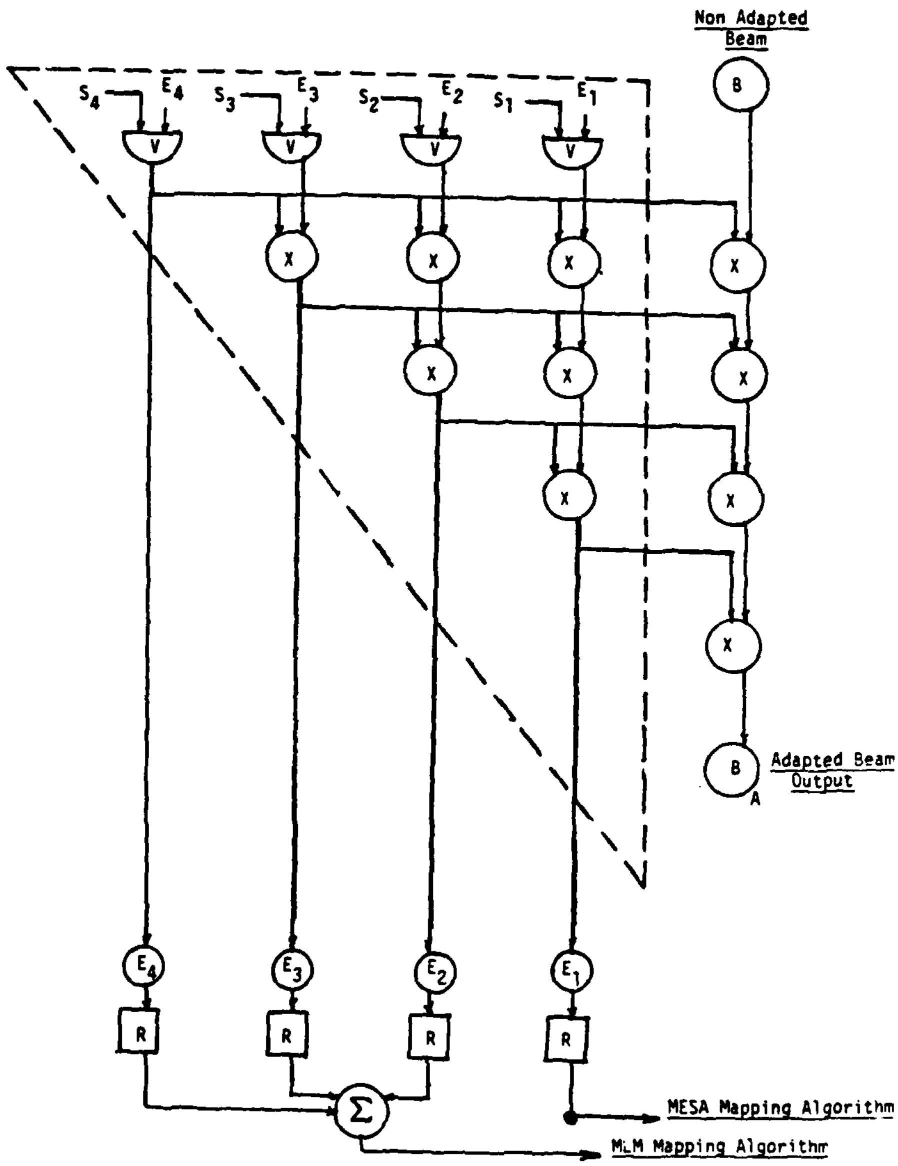
- Instantaneous Map (No Scanning Required)
- Near Optimum

TECHNIQUES

- MESA
- MLM
- WI
- AAR

- Thinned Array

SLIDE 15



COMBINED SLC AND MAPPING ALGORITHM

SLIDE 17

A DIGITAL ADAPTIVE ARRAY PROCESSING ALGORITHM FOR COMMUNICATIONS

IRVING REED: This talk is an outgrowth of an idea that Larry Brennan and I had when one time we looked at the JTIDS problem. About 3 and a half years ago we had a Naval Air Systems Command contract which was monitored by Jim Willis. Our primary work had been in adaptive arrays as applied to radar, but one year Jim Willis decided he should also be concerned about communications. So we looked at the various possible spread-spectrum communications applications of adaptive antennas. Since I knew a little about JTIDS at the time, I suggested we look at what could be done with such a system. What we did was to devise an adaptive algorithm scheme using antenna elements that in the presence of jamming and interference would automatically make synchronization and the selection of signals possible. At the outset I want to mention that the idea for this talk is in a paper published in the IEEE Trans. on Aerospace and Electronics systems. The title for this paper is, "An Adaptive Array Signal Processing Algorithm for Communications," Vol. ASAT, January 1982.

The important property of this nulling technique is that it does not require a knowledge of signal direction. At the same time the algorithm maximizes the probability of signal selection and/or synchronization. The implementation of this technique I won't explicitly talk about. John Bailey has already done this very well. Needless to say, such a system could utilize a digital processor system somewhat like JTIDS now is configured.

The present adaptive algorithm for communications is related to the standard adaptive array algorithm which utilizes a pilot signal. This is the technique which was devised by Widrow. It is the so-called LMS or adaptive Wiener filter algorithm. One starts with a set of pilot signals or signals one wants to correlate against. The best mean square filter weights are then obtained by a set of time correlations. In Figure 3 one has a box that might be a digital processor. In a communication system before synchronization one has two hypotheses. Either a signal is now arriving or one has noise only. There are N elements in the array and K input samples for each of these N elements.

In Figure 4 the signal transmitted is the sequence P_{ck} and S is the signal direction vector. For a linear array this direction steering vector is an array of phasors looking in the direction of the signal. However, this angular information is not required to be known in the criterion.

The idea of this algorithm is straightforward see Figure 5. It's an extension of a Wiener least squares estimation scheme to a multiple hypothesis criterion. For signal selection one selects one signal from among a set of L possible signals that have been transmitted. For synchronization one selects one waveform from among different phases of a header sequence. If one uses the appropriate header sequence, L different filters are needed to determine the time or phase that the sequence was transmitted.

The algorithm requires that the mean square error be found between a weight vector times X_k and the header sequence $P_{\ell k}$. Then one minimizes with respect to both W and ℓ. The minimization over the two parameters W and ℓ, is accomplished in two steps. First minimize with respect to W and then minimize with respect to ℓ.

As shown in Figure 5, the actual error that one observes for the ℓ^{th} signal

is essentially a quadratic form. M is the estimated covariance function and this other term XP_{ℓ}^{*} is the sampled correlation between the actual received data and the signal itself.

If one minimizes over ℓ one can see now that it would be nice to make this quadratic form independent of the signal structure. This is achieved by assuming the sequences $P_{\ell k}$ are mutually orthogonal. If these sequences are not actually orthogonal, the present analysis is still approximately correct. If the L signals, which can be either time shifts of a header sequence or the M signals in the communication process, are assumed to be orthogonal, the constant $P_{\ell}P_{\ell}^{*}$ becomes one, and one gets one minus a simple quadratic form to minimize with respect to ℓ. This is the same thing as a maximization of the quadratic form. To recapitulate in Figure 6 first form the covariance matrix, next estimate it's signal vector for all ℓ and then form the weights W_{ℓ} for all L. Finally form q_{ℓ} for all possible signals and then maximize with respect to ℓ.

To recapitulate again in Figure 7 so that everyone understands, compute q_{ℓ}, on up to q_{L}. To do this, compute the covariance matrix, form the correlation, and form the ℓ quadratic forms. Here ℓ is the number of codes to be tested, and N is the number of elements in the array. Finally K is the number of samples in the train that the algorithm observes.

Larry and I, after a great deal of work managed to find most of the probability distributions needed to analyze this problem, not perfectly, but almost. In Figure 9 we found by methods pioneered by Goodman for the complex Wishart distribution, the actual probability density as a function of ρ where ρ is the generalized SNR, K is the number of samples, and A^2 is the signal power. ρ is the so-called generalized SNR and M is the true covariance matrix of the noise.

The probability distribution can be represented as a confluent hypergeometric function. This fact was used by Martin Cohen for calculation purposes. For the incorrect codes one doesn't know the correct probability density. We only know the density when the incorrect signal equals zero. Or in other words, this density is accurate only for a small input SNR. However, assuming this to be true (see Figure 10) one can obtain an estimate of the probability of making a correct decision among the L possible codes.

I must apologize for some of the computer plots. This contract ended over 2 years ago, and the computations were done on a Pet computer after the contract ended. Martin Cohen, remarkably enough, wrote a plotter program for the Pet computer. In Figure 11 are shown probability densities as a function of ρ where ρ is the generalized SNR. For ρ equal to zero this is a symmetric distribution, and as ρ, the generalized output SNR gets larger, these distributions shift to the right. Since ρ lies between 0 and 1, they get closer and closer to 1.

The integral in Figure 10 I like to call the N-horse integral. A similar integral can be used to compute the probability that any one horse will beat all the other horses of horse race. This integral is presented in Figure 11 for various values of ρ. ρ goes from 0 to 64. The number of elements is N=8. As you can see, with the output SNR and the number of alternative symbols sufficiently large one gets a relatively small probability of error.

Actually this doesn't tell the whole story, and for this reason I made some more viewgraphs last night. However, before presenting them, note that from Figure 12 there are still a number of unsolved problems. We don't know exactly the probability distribution of the output

which includes precisely the effects of signals on the covariance matrix. This is a problem we have so far not been able to solve. However, we did find the characteristic function for such a probability distribution. There is also a problem associated with the sampling misalignment time which hasn't been analyzed. Someday, if there is sufficient interest, we will have to carry out a simulation of this problem.

Another way of looking at this problem is to find an improvement factor. Suppose in Figure 13 that P_0 is the receiver noise power and P_1 is the jammer noise power. Then for jamming, the SNR before adaptation P_1 is much greater than P_0. P_1 could be quite a bit greater, like 30 dB greater, and the SNR at the output of a standard 1 element receiver would have this SNR. The SNR with adaptation would be this generalized SNR which is called ρ in our paper. In this formula I mean by S_0 the steering vector in the direction of the signal source. For a linear array, as I mentioned in a previous slide, these are the phasors for a linear array pointed in direction θ_0. I define what might be called the adaptive gain G which is the SNR, adapt, over SNR, not-adapt. With this definition of G one can make a performance improvement analysis. For the signal jammer case (see Figure 14) the covariance function is the receiver noise times an identity matrix plus the jammer power times a steering vector S_1 which is in the direction of the jammer. In this particular case, $S_1S_1^*$ is a rank 1 matrix. One can actually invert this matrix very simply and obtain M^{-1} in terms of P_0,S_1, N and $\mu=P_1/P_0$, the power ratio. With this in the formula for G one obtains G upon a division by S/N, non-adapt. Finally this gain function is N times the ratio of powers times the magnitude of $1-\beta^2$ where β is the magnitude of a normalized inner product between the two steering vectors, S_0 and S_1. Here vector S_0 points in the direction of the signal and S_1 points in the direction of the jammer.

With a little effort, and this is not exact, one can show for a linear array see Figure 15 that β^2 is approximately a sync function where δ is the difference in the sines of these two angles. β^2 peaks up and has a "beam width", which is a function of the difference in these two angles. If one ignores contributions outside this main beam and assume δ has the uniform distribution, the average of β^2 is roughly 1/4. Thus approximately the average gain is in the order of approximately 3/4 N times P_1/P_0. An example of how much gain one can get, assume 4 elements, and $P_1/P_0=10^2$. The average gain of such a system is roughly 24 dB.

A lot more work is needed on this adaptive criterion. There is certainly a need for a simulation of this concept. We've found some of the analytical problems associated with this problem to be hard. Though the distributions we've found are clever, it's about the end of the line of where analysis can go. That's about all I have time to say for the present.

WEBER: Any questions or comments?

SIMON: Could I ask you to go back to Figure 13?

REED: As I said, I did this only yesterday in consultation with John Bailey so I can't give more details at this time.

SIMON: The result that you have there, just looking at it quickly seems to be very similar to the result Applebaum originally got in his paper for his algorithm. This seems to be a generic problem which I haven't seen solved. At least I've played with it and haven't been able to do much with it. How do you do these problems when you have more than one jammer. Basically, how do you invert a

matrix like that when you have more than one jammer?

REED: As I say, you just add more terms, like this term here. There is a standard technique for inverting such matrices

SIMON: 2 jammers at different powers at different locations?

REED: Right. Actually there's a paper by Zaum where he does exactly that. He does the explicit matrix inversion.

SIMON: I'd like to see that.

REED: It's in the IEEE AES Transactions, I think. I don't remember the exact reference. But it can be done by a similar means. It's an extension, which makes use of the fact that you have rank 1 matrices.

A DIGITAL ADAPTIVE ARRAY PROCESSING ALGORITHM FOR COMMUNICATIONS

- NULLS JAMMERS AND INTERFERENCE
- DOES NOT REQUIRE KNOWLEDGE OF SIGNAL DIRECTION
- MAXIMIZES PROBABILITY OF CORRECT SYMBOL SELECTION OR SYNCHRONIZATION
- DIGITAL PROCESSOR

FIGURE 1

DIGITAL ADAPTIVE ARRAY - SAMPLE COVARIANCE MATRIX ALGORITHM

$$X_k = \begin{pmatrix} x_{1k} \\ x_{2k} \\ x_{nk} \end{pmatrix} = \text{column vector of } k^{th} \text{ sample of element outputs}$$

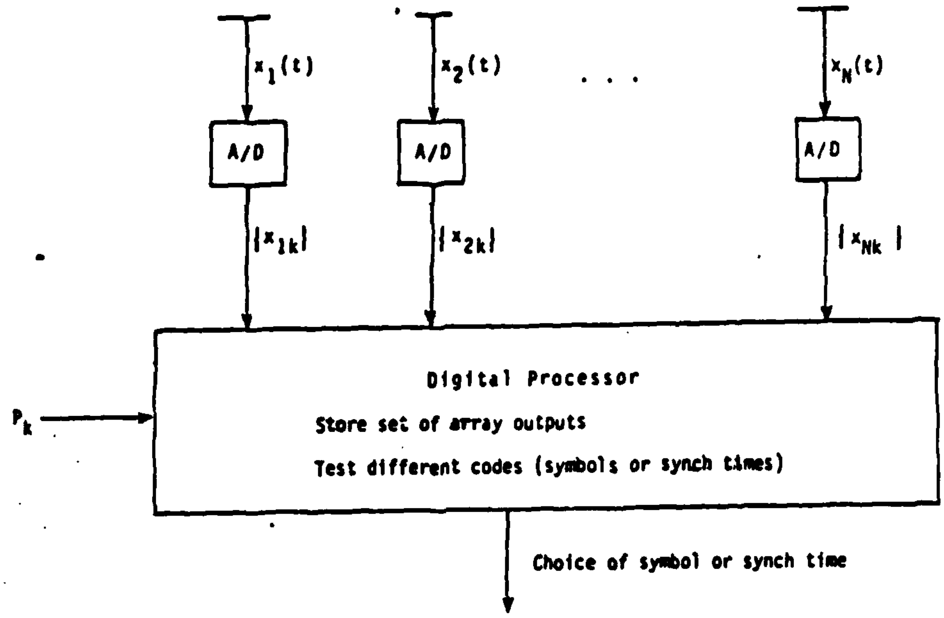
$$\hat{M} = \frac{1}{K} \sum_{k=1}^{K} X_k X_k^* = \text{sample covariance matrix}$$

$$\overline{\hat{Xp^*}} = \frac{1}{K} \sum_{k=1}^{K} X_k p_k^* \;\; ; \;\; p_k = k^{th} \text{ sample of pilot signal}$$

$$\hat{W} = \hat{M}^{-1} \, \overline{Xp^*}$$

$$Z = \hat{W}^* X = \text{array output}$$

FIGURE 2



ADAPTIVE ARRAY PROCESSOR FOR COMMUNICATIONS

FIGURE 3

During one symbol interval:

X_k = column vector of array outputs $(k=1,2,\ldots K)$

$$X_k = N_k + aSp_{ck}$$

N_k = column vector of noise components (complex)

a = unknown signal amplitude (complex)

S = signal direction vector

p_{ck} = k^{th} sample of correct (transmitted) code

For a linear array with element spacing d

$$S = \begin{pmatrix} 1 \\ e^{i \frac{2\pi d}{\lambda} \sin\theta} \\ \vdots \\ e^{i \frac{2\pi d}{\lambda} (N-1)\sin\theta} \end{pmatrix}$$

FIGURE 4

Test L different codes, $\ell = 1,2,\ldots,L$

$$P_1 = p_{11}, p_{12}, \ldots, p_{1k}$$

$$P_L = p_{L1}, p_{L2}, \ldots, p_{Lk}$$

Find weights for ℓ^{th} code such that

$$\epsilon_\ell(w) = \frac{1}{K}\sum_{k=1}^{K} |W^* X_k - p_{\ell k}|^2 \text{ is minimum}$$

$$W_\ell = \hat{M}^{-1}\,\overline{Xp_\ell^*} \quad \text{minimizes } \epsilon_\ell$$

$$\epsilon_\ell(W_\ell) = \overline{p_\ell p_\ell^*} - \overline{p_\ell X^*}\,\hat{M}^{-1}\,\overline{Xp_\ell^*}$$

where $$\hat{M} = \frac{1}{K}\sum_{k=1}^{K} X_k X_k^*$$

$$\overline{Xp_\ell^*} = \frac{1}{K}\sum_{k=1}^{K} X_k p_{\ell k}^*$$

FIGURE 5

Form sample covariance matrix, $\hat{M}$, and estimated signal vector $\overline{Xp_\ell^*}$

$$\hat{M} = \frac{1}{K}\sum_{k=1}^{K} X_k X_k^* \quad , \text{ same for all codes, } \ell$$

$$\overline{Xp_\ell^*} = \frac{1}{K}\sum_{k=1}^{K} X_k p_\ell^* \qquad \ell = 1,2,\ldots,L$$

Weights for code ℓ, W_ℓ

$$W_\ell = \hat{M}^{-1}\,\overline{Xp_\ell^*}$$

Output for code ℓ

$$W_\ell^* X = \left(\overline{p_\ell X^*}\right)\hat{M}^{-1} X$$

Average signal power in output q_ℓ

$$q_\ell = \overline{p_\ell^*(W^* X)} = \left(\overline{p_\ell X^*}\right)\hat{M}^{-1}\left(\overline{Xp_\ell^*}\right)$$

FIGURE 6

$$\epsilon_\ell(W_\ell) = \overline{p_\ell p_\ell^*} - \left(\overline{p_\ell X^*}\right)\hat{M}^{-1}\left(\overline{X p_\ell^*}\right)$$

Let p (=1,2,...L) be pure phase codes and orthogonal

$$\overline{p_\ell p_\ell^*} = 1$$

$$\overline{p_m p_\ell^*} = 0 \quad m \neq \ell$$

For each code $p_1, p_2, \ldots P_L$, find W_ℓ which minimizes ϵ_ℓ

Select as correct code the p_ℓ for which ϵ_ℓ is minimum where

$$\epsilon_\ell = 1 - \left(\overline{p_\ell X^*}\right)\hat{M}^{-1}\left(\overline{X p_\ell^*}\right)$$

or for which

$$q_\ell = \left(\overline{p_\ell X^*}\right)\hat{M}^{-1}\left(\overline{X_{p\ell}^*}\right) \text{ is maximum}$$

FIGURE 7

To compute $q_1, q_2, \ldots q_L$:

Form and invert $\hat{M}$ once

$$\hat{M} = \frac{1}{K}\sum_{k=1}^{K} X_k X_k^*$$

$\hat{M}$ is hermitian NxN matrix

Form L $\overline{X p_\ell^*}$ column vectors

Form L quadratic forms, q_ℓ

$$q_\ell = \left(\overline{p_\ell X^*}\right)\hat{M}^{-1}\left(\overline{X p_\ell^*}\right)$$

Select ℓ with largest q_ℓ as correct output

L = number of codes tested

N = number of elements in array

K = number of samples

FIGURE 8

For correct code, probability density is:

$$P_1(q) = e^{-q}\sum_{m=0}^{\infty}\frac{(m+K-1)!\,\rho^m q^{N+m-1}(1-q)^{K-N+1}}{m!\,(N+m-1)!\,(K-N-1)!} \quad , \quad 0 \le q \le 1$$

$\rho = K|a|^2 S^* M^{-1} S$

K = number of samples

$|a|$ = signal amplitude

S^* = signal steering vector

M = true covariance matrix of noise

$q = \overline{P_\ell X^*}\, \hat{M}^{-1}\, \overline{X P_\ell^*}$ contains a signal component

$X = N + aS\rho_\ell$

$\overline{X P_\ell^*} = aS \neq 0$ (true mean)

FIGURE 9

For incorrect codes, probability density is:

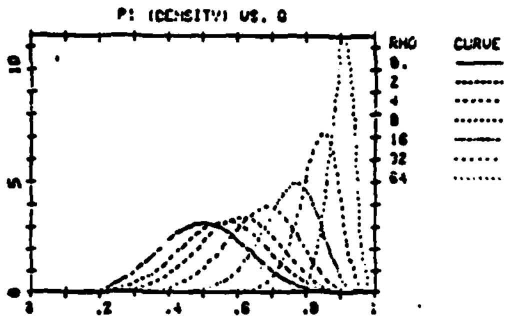
$$Po(q) = \frac{(K-1)!\,q^{N-1}(1-q)^{K-N+1}}{(K-N-1)!\,(N-1)!} \quad , \quad 0 \le q \le 1$$

q contains no signal component, $\overline{X P_\ell^*} = 0$ (true mean)

Probability of correct decision among L codes is:

$$P_c = \int_0^1 P_1(q)\left(\int_0^q P_0(s)\,ds\right)^{L-1} dq$$

FIGURE 10



PE FOR K = 16, N = 8, INTEGRAL POINT COUNT = 40

M	ALPHA = 0.	.125	.25	.5	1	2	4
	RHO = 0.	2	4	8	16	32	64
2	.500000	.377240	.282280	.158744	.040615	.006317	.000184
4	.750000	.625018	.510005	.323509	.122773	.010103	.000733
8	.875000	.781872	.681540	.480859	.229625	.038750	.001813
16	.937500	.875886	.800484	.631230	.336450	.072317	.003012

MAX ACCURACY POSSIBLE = 1.17274803E-04

Probabilities of Error in Symbol Selection

K = No. Samples = 16
N = No. Elements = 8
M = No. Alternative Symbols
RHO = S/N Ratio

FIGURE 11

UNSOLVED THEORETICAL PROBLEMS RELATING TO THE DIGITAL ADAPTIVE ARRAY ALGORITHM

- PROBABILITY DISTRIBUTION OF OUTPUT (OR q) FOR WRONG HYPOTHESIS, INCLUDING EFFECT OF SIGNAL ON COVARIANCE MATRIX
- SAMPLING MISALIGNMENT IN TIME
- EFFECT OF CORRELATION BETWEEN DIFFERENT OUTPUTS (OR q)

FIGURE 12

ADAPTIVE GAIN

- S/N WITHOUT ADAPTATION (SINGLE ELEMENT):

$$(S/N)_{\overline{ADAPT}} = \frac{K|A|^2}{P_0+P_1} \approx \frac{K|A|^2}{P_1}$$

WHERE P_0 IS RECEIVER NOISE POWER

P_1 IS JAMMER NOISE POWER

K IS NUMBER OF WAVEFORM SAMPLES

$|A|^2$ IS SIGNAL POWER/SAMPLE

- S/N WITH ADAPTATION (N ELEMENT ARRAY):

$$S/N)_{ADAPT} = K|A|^2 \, S_0^* \, M \, S_0 \equiv \rho$$

WHERE M IS COVARIANCE MATRIX FOR RECEIVER AND JAMMING NOISE,

S_0 IS STEERING VECTOR TO SIGNAL SOURCE

- FOR LINEAR ARRAY

$$S_0 = \left| 1, e^{\frac{2\pi i D \sin \theta_0}{\lambda}}, \ldots e^{\frac{2\pi i D(N-1) \sin \theta_0}{\lambda}} \right|^2$$

- ADAPTIVE GAIN = G

$$= (S/N)_{ADAPT} \Big/ (S/N)_{\overline{ADAPT}}$$

FIGURE 13

G FOR SINGLE JAMMER CASE

- COVARIANCE MATRIX:

$$M = P_0 I + P_1 S_1 S_1^*$$

WHERE I IS IDENTITY MATRIX

S_1 IS STEERING VECTOR TO JAMMER

- INVERSE OF M:

$$M^{-1} = \frac{1}{P_0}\left[I - \frac{\mu S_1 S_1^*}{1 + N\mu}\right]$$

WHERE $\mu = P_1/P_0$

- $\rho = (S/N)_{ADAPT}$,

$$= \frac{K|A|^2}{P_0}\left[N - \frac{\mu|S_0^* S_1|^2}{1 + \mu N}\right]$$

- FOR $P_1 >> P_0$ GAIN IS:

$$G = N\mu\left|1 - \beta^2\right|$$

WHERE β IS MAGNITUDE OF NORMALIZED INNER PRODUCT OF VECTORS S_0 AND S_1.

FIGURE 14

- FOR LINEAR ARRAY

$$\beta^2 = \frac{1}{N^2}\,\frac{\sin^2\left(\frac{\pi D N \Delta}{\lambda}\right)}{\sin^2\left(\frac{\pi D \Delta}{\lambda}\right)}$$

$$\approx \operatorname{sinc}^2\left(\frac{\pi D N \Delta}{\lambda}\right)$$

WHERE $\Delta = \sin\theta_1 - \sin\theta_0$

- ASSUME Δ HAS UNIFORM DISTRIBUTION

$$\text{Ave}\,(\beta^2) \approx \frac{1}{4}$$

- AVERAGE ADAPTIVE GAIN

$$G_{AVG} = \frac{3}{4}\,N\;P_1/P_0$$

- EXAMPLE $N = 4,\; P_1/P_0 = 10^2$

$$G_{AVG} = 3 \times 10^2 \approx 24\text{ dB}$$

FIGURE 15

GENERAL DISCUSSION

BEDROSIAN: I'm unencumbered by the need to do any work in this area. In fact I'm further unencumbered by not really being familiar enough with the material so it becomes difficult for someone like myself to ask questions for fear that you're going to say something silly or trivial. I had a thought yesterday and I chatted with Bob Price about it and he encouraged me to put it before the group. It came about in the discussion I think I heard yesterday about what do you do when you get this variety of jammers of different types with whom you have to cope ? You've selected a coding scheme and then as I hear the speakers there are a variety of ways you can decode, there are a variety of ways you can process the signal, some are good against one type of jamming, some are good against another type of jamming, and many of the talks seem to hinge on what you should do. And while you can calculate everything you need to calculate, my question would become, as an outsider, "Why not do them all?" Process every different way you can think of. You've collected the received data. From my superficial view of it, I'd say that that might not be a bad idea except that when you are done, I can appreciate that having processed every different way you could think of, you still may not know which one of these is the best processing method. And after a while I remembered error-correction, so I'll ask this question. Isn't it possible to code in such a way that one can look at the variously processed outputs and derive some measure as to which of these is the best? Then you can use that one to the exclusion of others. What I don't know are the details about how to do this. Bob had some thoughts. Do you want to add those now ?

BOB PRICE: I thought it was a very nice question he posed to me that hadn't occurred to me. Although I had asked about the possibility of ARQ where you ask the transmitter to repeat. I think Seymour felt that maybe the return link was not reliable enough to do this, so this sounded analogous to that in that the receiver with the suitable recording, say, and then massaging over and over again by different processing algorithms. This would essentially achieve ARQ type operation by ultimately hitting the right combination of nulls and so forth and finally receiving the message. So Ed asked me what kind of measure could tell you when you were on the right one And it occurred to me that error-detection is duck soup, I mean we all slave over error-correction but error-detection to a level where the undetected error probability is essentially zero or 10^{-64}. Something like that is duck soup. So I suggested to Ed that that could be his measure that you would simply look at the final result and see if it had any residual errors in it or not. Of course if they all had residual errors you would have failed but if there were only one of them, or any set of them happened to have an indication that there were no residual errors then you would have accomplished your mission.

Ed was suggesting sheer parallelism. I was trying to save money by hoping that there would be enough time to repeat and try the reprocessing and reprocessing. And I'm glad you asked the question about how do we record it because that uses something like optical link communications which hasn't come up in this meeting at all. And it might be worthy of study. There used to be wideband recording for radar data. Some one might look at that as a resort, that would be expensive, and wouldn't use magnetic recording. So that was a thought.

And I'd like to also take this opportunity to throw out a gem I learned

about this morning. And it's completely non-military, but those of you in the business might like to know this that City Bank in New York has under contract a company which I've never heard of called Equatorial Radio Inc. They are looking into the possibilities of spread spectrum communications for local looping from satellites to small dishes on the ground, presuming there's no jamming threat there. City Bank has a policy of never using the telephone company if they can help it. They feel that spread spectrum has some answers for them in the semi-cryptology area. It's very preliminary and they've cautioned me not to go gung-ho on this the way Hewlett Packard has, but it was interesting for me to know that.

COMMENT: A representative from Equatorial was at that Long Island spread spectrum conference a couple of weeks ago and they are doing more than looking into it. They are making a gross of 95 million this past year or something like that. I may even be remembering this wrongly, but they made money this past year and they use a 2-foot dish which you can buy for $2,500. I have a brochure on this in my briefcase.

GAYLORD HUTH: City Bank thinks this is a good investment of their venture capital. Something might come out of it. One other thing that wasn't mentioned in the conference. I may have been out of the room. It was about coherent frequency hopping and coherent frequency synthesizers. They are very expensive but is there any reason to push them along the way we might push recording systems along ?

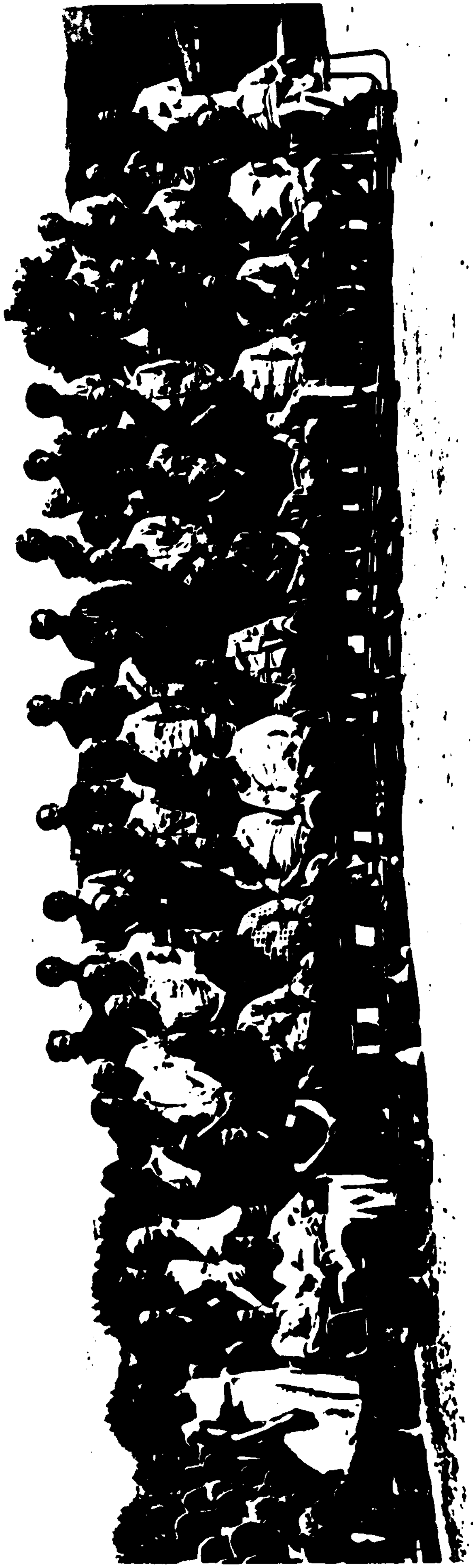
JOHN CAFARELLA: One of the reasons you might go to frequency hopping for very light spreading as opposed to a direct sequence is that you can cover that large bandwidth with cheaper components. Aside from the fact that you are only covering the band in small chunks, the post-processor circuitry is simpler. You tend not to have really the whole front end of the receiver be so tightly controlled across the entire band that you might use it coherently. Hence if you try to do coherent frequency hopping, you might have a serious problem in terms of the cost of the system.

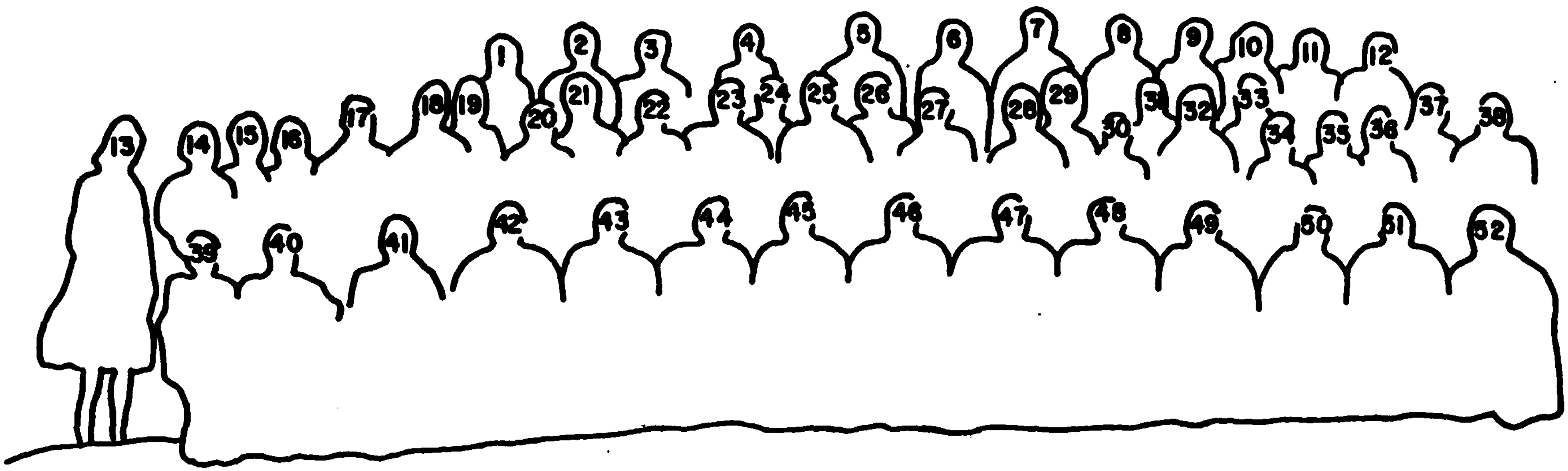
PRICE: You could have a coherent frequency hopper on the transmitter. If the cost of coherent frequency hopping could come down, the receiver could receive the signal and ignore the coherence if it wanted to but in some applications for super fine ranging it might be able to take advantage of the coherence. So I guess my argument is that coherent frequency hopping can't hurt and might help.

PURSLEY: I'd like to make a few comments on the parallelism. I've actually been studying this for Reed-Solomon codes where the simplest scheme is to simultaneously try to use channel side information to make erasures and feed it into an erasure correcting algorithm at the same time. At the same time you can make hard-decisions on the M-ary symbols by going to an error-correction algorithm. Now one advantage of Reed-Solomon codes is that with high probability if there is an error-rate that will not decode, it will not map into another code word. So you don't need the error-detection, it's automatically built in and it works very well. I've looked at the performance of that kind of system for both jamming and also for multiple access interference and have some papers written on that subject. But I think much more can be done, and I think it is the right way to go. The adaptivity is put at the receiver, who knows what's going on. The transmitter sometimes has difficulty knowing when its being jammed and when it isn't. However, the receiver can sit there and do anything he wants and process this with lots of parallelism. In many cases the

receiver will know the right answer because the decoder will tell him that it's hung up, I can't decode, so he forgets that word.

WEBER: Anyone else ? O.K. Thank you for attending.





1.	Tsun-Yee Yan	14.	Norman Krasner	27.	Donald Leimer	40.	Barry Leiner
2.	Joseph Yuen	15.	John Olsen	28.	Elwyn Berlekamp	41.	John Silvester
3.	Paul Feintuch	16.	Robert Travis	29.	Marvin Simon	42.	Charles Weber
4.	Laurence Milstein	17.	Peter Pawlowski	30.	Victor Li	43.	Lloyd Welch
5.	William Sander	18.	Michael Pursley	31.	Robert McEliece	44.	Robert Scholtz
6.	Marlin Ristenbatt	19.	Gaylord Huth	32.	Lawrence Rauch	45.	John Wozencraft
7.	Harold Fredricksen	20.	Seymour Stein	33.	Charles Graff	46.	James DuPree
8.	John Dillon	21.	Rodger Ziemer	34.	Lowell Rosen	47.	William Floyd
9.	Andreas Polydoros	22.	Bart Rice	35.	Jeff Wieselthier	48.	Irving Reed
10.	Jim Omura	23.	Fouad Tobagi	36.	Kesh Bakhru	49.	Robert Price
11.	Ronald Mathis	24.	Joseph Odenwalder	37.	Barney Reiffen	50.	Edward Posner
12.	Roger Peterson	25.	Robert Dinger	38.	Pravin Jain	51.	John Cafarella
13.	Mary Ann Kiefer	26.	Jurgen Gobien	39.	Paul Sass	52.	Edward Bedrosian

Not pictured: Barry Levitt, John Bailey

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